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Microwaves & RF

News
Generators deliver vector modulation

Design Feature
Tracking advances in VCO technology

Product Technology
UWB ICs send 100-Mb/s wireless data

YIGs Tune High-Speed MM-Wave Radios

Frequency Generation/Control Issue

MILLIMETER WAVE
MIXER ASSEMBLIES

From Stock

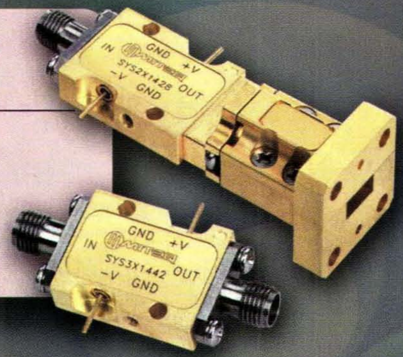
MIXERS

Model Number	Frequency (GHz)			LO Power (dBm)	Conversion Loss (dB Typ.)	LO-RF Isolation (dB, Typ.)
	RF	LO	IF			
TB0440LW1	4–40	4–42	.5–20	10–15	10	20
DB0440LW1	4–40	4–40	DC–2	10–15	9	25
SBE0440LW1	4–40	2–20**	DC–1.5	10–15	10	20
IR2640L17*	26–40	26–40	Note 1	15	10	15
M2640W1	26–40	26–40	DC–12	10–12	10	20
TB2640LW1	26–40	26–40	.5–20	10–15	10	20

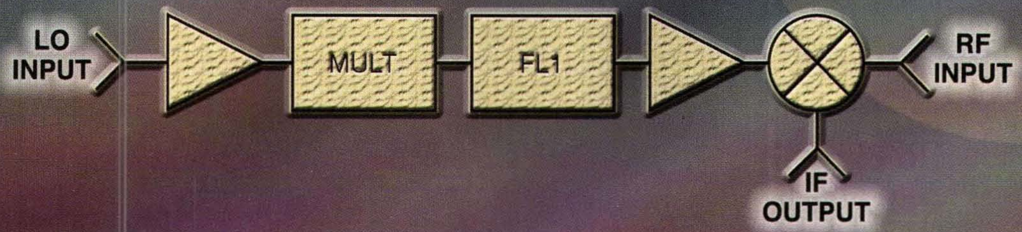
* Image Rejection typically 15 dB. ** Sub Harmonic
Note 1: IF Option A: 20–40 MHz, B: 40–80 MHz, C: 100–200 MHz, Q: DC–1000 MHz

MULTIPLIERS

Model Number	Frequency (GHz)		Input Power (dBm)	Output Power (dBm, Typ.)	Fundamental Leakage (dBc, Typ.)
	Input	Output			
SYS2X1428	14	28	+12	+12	-50
SYS2X1734	16–17.5	32–35	+12	+12	-50
SYS3X1442	14	42	+12	+12	-50
SYS4X1146	11	46	+12	+15	-60
SYS2X2040	10–20	20–40	+12	+15	-15
TD0040LA2	2–20	4–40	+10	-5	-20



MIXER/MULTIPLIER ASSEMBLIES



Model Number	Frequency (GHz)			LO Power (dBm)	Conversion Loss (dB, Typ.)	Input IP^3 (dBm, Typ.)	Fundamental LO-RF Isolation (dB, Typ.)
	RF	LO	IF				
SYSMM2X2335	23.67–35.33	11.385–17.665	.04–.230	13–15	12	+15	50
SYSMM3X2640	26.5–40	8.8–13.3	DC–.5	10	10	+15	40

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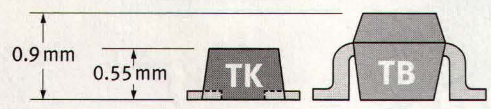
INTRODUCING THE WORLD'S SMALLEST GaAs MMIC SWITCH



UPG2008TK *SOT-363*

Meet the UPG2008TK. It's footprint is less than half that of a standard SOT-363 switch. Plus its leads are flat and recessed into the base of its package, giving it that super low profile your miniaturized portable products demand. Optimized for transmit/receive switching, antenna diversity switching, and power level control, the UPG2008TK is the latest in our growing family of GaAs MMIC SPDT switches.

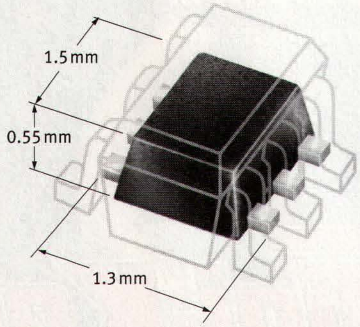
See them all on our web site at:

www.cel.com/spdt.asp

GaAs MMIC SPDT Switches

Part Number	Insertion Loss @ 1.0 GHz	P_{IN} Power Handling	Control Voltage	Package	100K Price	Description
UPG2008TK	0.4 dB	+25 dBm @ 1.0 dB	2.8 V	TK	45¢	*World's Smallest*
UPG2009TB	0.25 dB	+34 dBm @ 0.1 dB	2.8 V	TB	78¢	*High Power, No Compromises*
UPG2006TB	0.35 dB	+20 dBm @ 1.0 dB	1.8 V	TB	54¢	*Low Voltage, Great Specs*
UPG158TB	0.3 dB	+25 dBm @ 0.1 dB	3 V	TB	39¢	*Good Specs, Great Price*
UPG152TA	0.4 dB	+30 dBm @ 1.0 dB	3 V	TA	29¢	*Low Cost 3V Switch*

Applications for CEL's NEC switches include: Bluetooth, PCMCIA Cards, Cordless Phones, Digital Cellular, PCS, WLAN, and WLL.



NEC's new TK package has a 50% smaller footprint than the TB (SOT-363) package switches.

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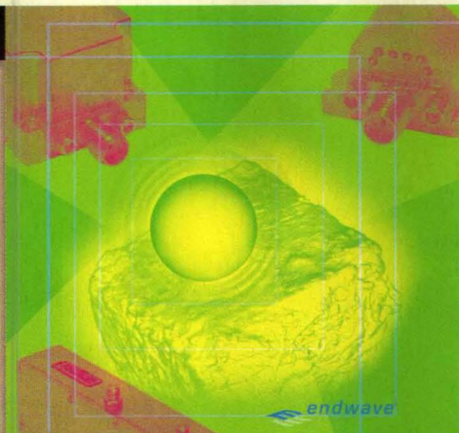

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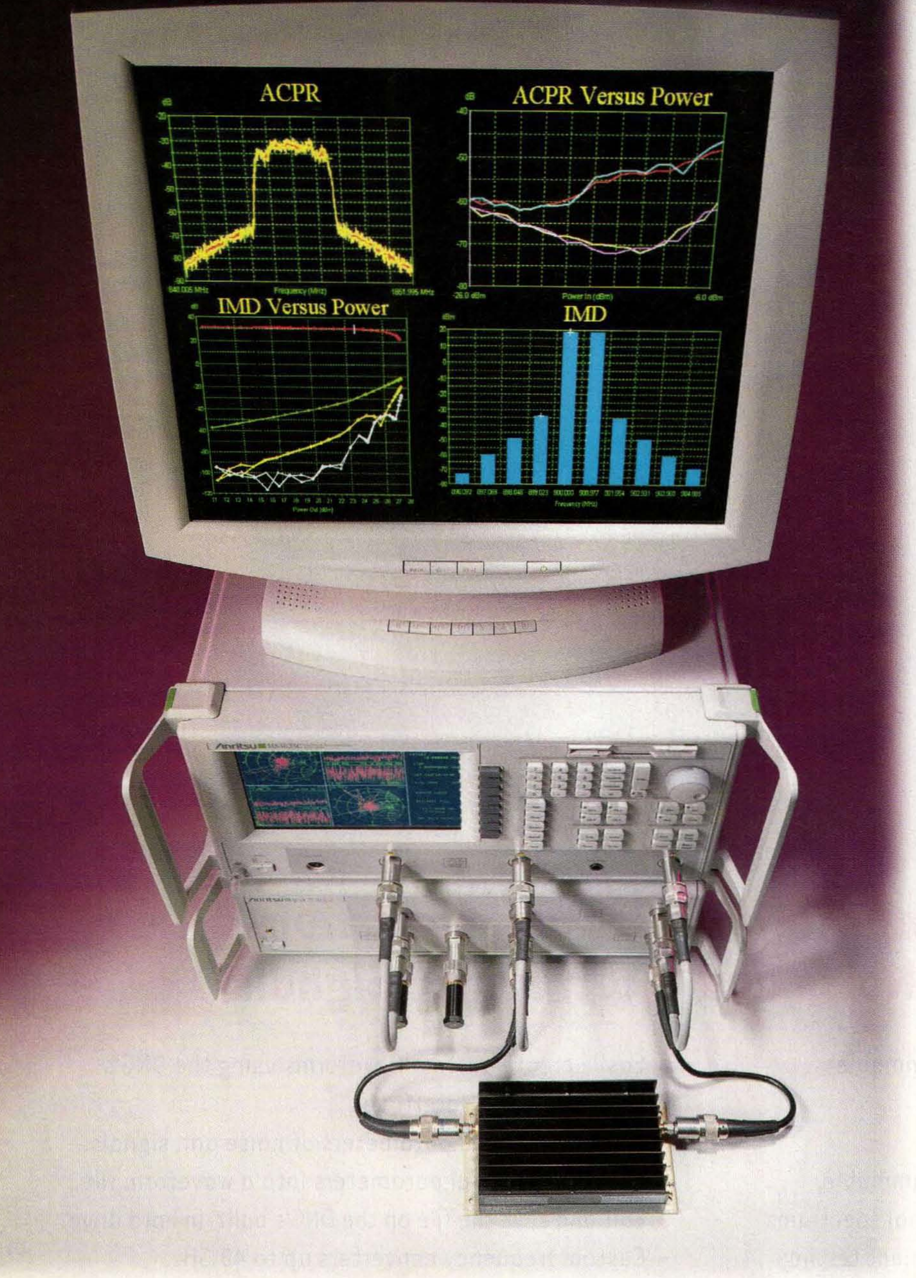

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Missing Information

▶▶THERE WERE SEVERAL errors that appeared in my article entitled "Scanning Technique Sorts Skin Permittivity" (*Microwaves & RF*, May 2002, pp. 106-114):

- In equations (2) and (3): φ should be replaced by $\approx$.
- In equation (3): λ_r should be replaced by μ_r.
- In equations (7), (8), and (9): κ_s should be replaced by λ_s.
- In equations (12) and (13): o should be replaced by π.
- In equation (15): $\}$ should be replaced by $>$.
- Figs. 7 (a) and (b) are missing from the article.

Before the last paragraph of the article, the following text is missing:

This signal goes to the first balance mixture and it is mixed with $f_o \pm f_m$. The output of the first balanced mixture ($f_i \pm f_m$) is connected to the second balanced mixture with an IF amplifier. In this mixture, $f_i \pm f_m$ and f_i are mixed and an output (f_m) goes to the PSD. The output of the PSD is connected to the y-axis of a plotter. The output of the vibrator is connected to the y-axis of a plotter with an x-translator. The output of the PSD (y-axis) is calibrated in ϵ_{rd} and the x-axis is calibrated with the sample size of the dielectric/biological samples under test.

A 1-W, 10-GHz source is used to illuminate the 300-μm aperture in the microstrip resonator. The samples are positioned at a mean position of approximately 50 μm away from the microstrip ground plane and they are vibrated (at 300 Hz) at an amplitude of approximately ±30 μm. The samples are moved beneath the aperture at a constant speed of approximately 2.5 mm/min.

Figure 7a shows a scan from the edge of a polystyrene sample to plexiglass sample, where it can be seen that a variation of only 2.5 percent in relative permittivity is easily detected. From the figure, it can be concluded that this instrument is very sensitive to detect variation in dielectric property of material. **Figure 7b** shows a plot of a man-made biological sample of a large variation in terms of relative permittivity.

One sample of a biological sample is normal skin and cancerous tumor cells (connected to the skin). Relative permittivity of the skin and the cancerous tumor are 30 and 54[3,15-16], respectively. In Figs. 7a and b, the theoretical and the experimental curves are shown by dotted and full lines, respectively. There is a very small deviation between the theoretical and the experimental plots due to measurement error.

A. Kumar
President
AK Electromagnetique, Inc.


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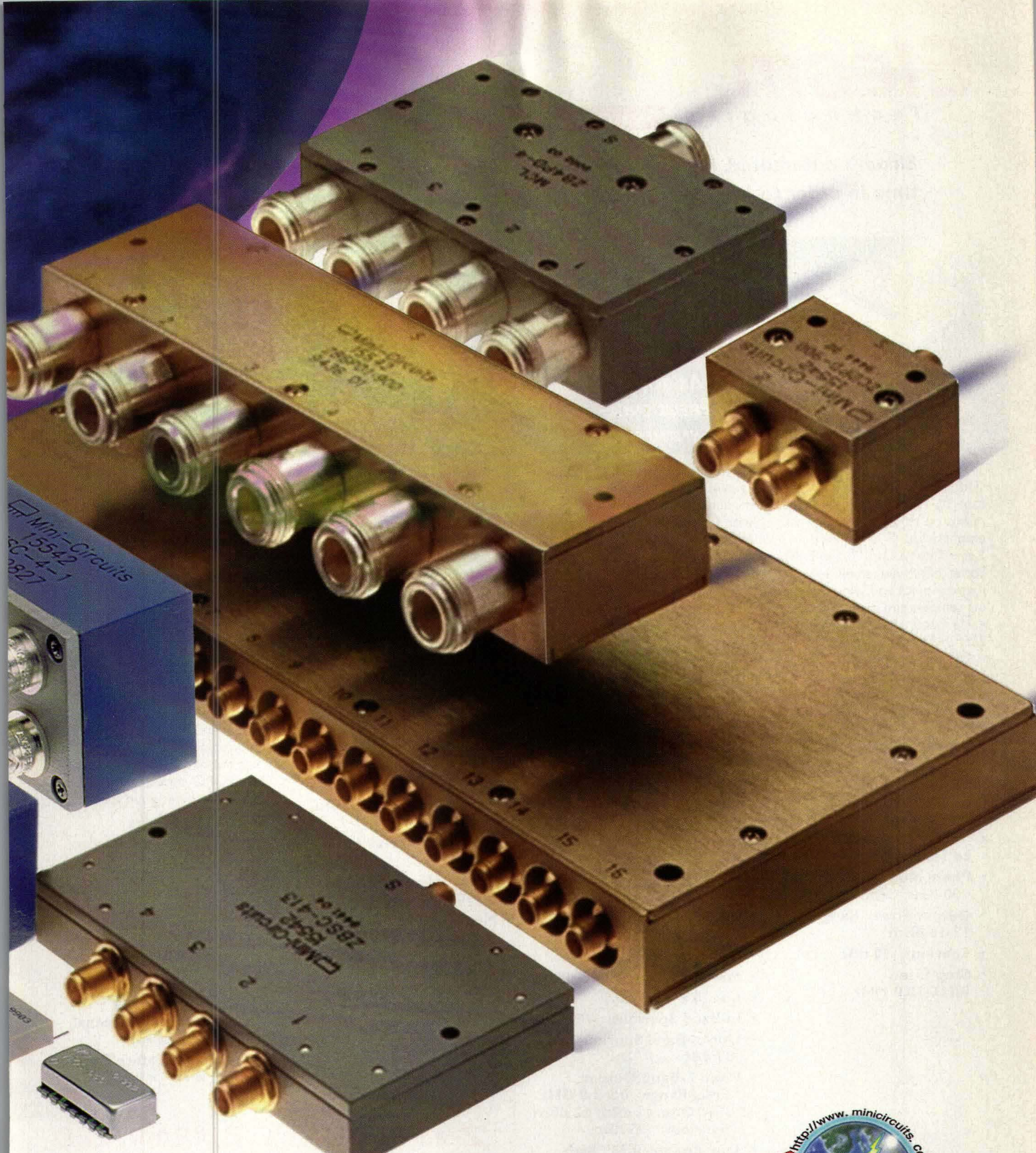
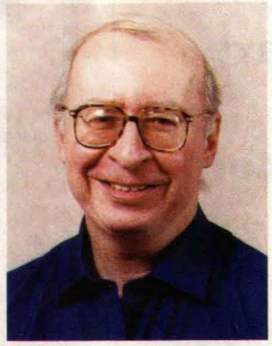


Searching For That Tactical Edge



Warfare and tactical methods are changing almost day by day. The days of conventional warfare where troops from either sides of a conflict lined up on opposite ends of a battlefield are long gone. World Wars I and II and the fruits of the Industrial Revolution introduced advanced technologies to the battlefield on a global scale. Latter-day conflicts such as Operation Desert Storm demonstrated our shift toward a highly mechanized military approach with sophisticated air-strike capabilities.

Operation Desert Storm demonstrated our shift toward a highly mechanized military approach with sophisticated air-strike capabilities.

The threat of large-scale conflicts may always be with us, and today's military forces must be prepared to mobilize at a moment's notice, similar to a modern counterpart to the "Minute Men" of the American Revolutionary War. But unlike that war, where our forces were defending their own turf, modern conflicts can take place almost anywhere on the planet.

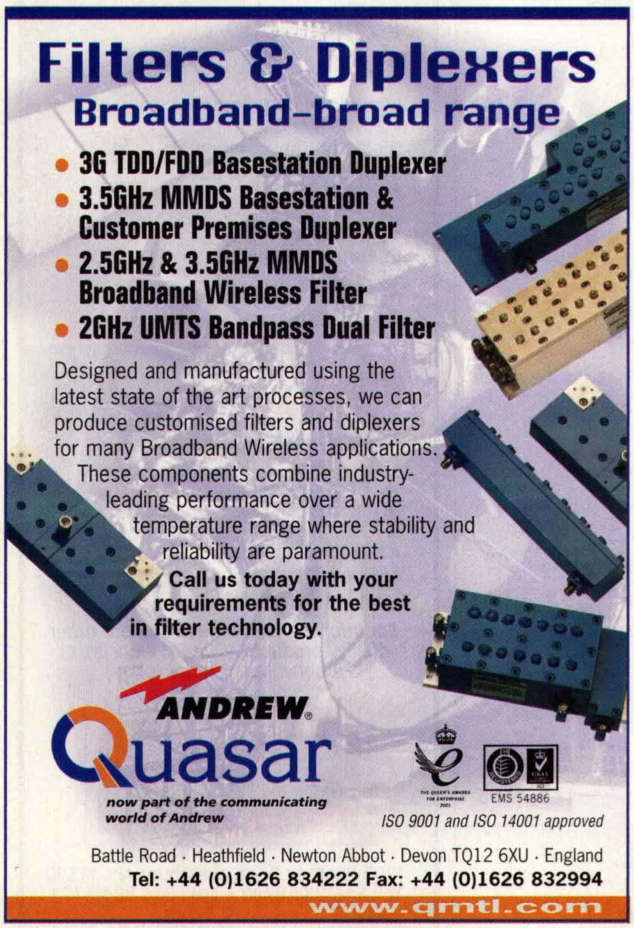
Ideally, war will one day cease to exist. But human nature being what it is, that day may not come soon. If modern military electronics technology does try to make warfare more "efficient," it also strives to ultimately save lives by bringing a swift end to a conflict. For example, the Boeing Missile Defense Systems Division (Canoga Park, CA) is currently negotiating with the US Special Operations Command (MacDill AFB, FL), which is acting on behalf of the US Department of Defense (DoD), to demonstrate the feasibility of using tactical lasers in place of missiles and munitions.

Technology for unmanned airborne vehicles (UAVs) continues to move ahead. Companies such as BAI Aerosystems (Easton, MD) have succeeded in developing short-range "disposable" UAVs and dense tracking and video circuitry for a variety of applications, including reconnaissance, surveillance, targeting, and remote sensing. In fact, the company's Exdrone UAV was used by the US Marines during Operation Desert Storm.

Helping military technologies to advance through the interaction of engineers is part of the charter of the Military Electronics Show (www.mes2002.com), scheduled for September 24-25, 2002 in the Baltimore Convention Center. Speakers include Dick Bernstein, President and CEO of BAI Aerosystems on the company's UAV technology, and Uri Yaniv, President and Chief Engineer of Elcom Technologies (Rockleigh, NJ) on his company's fast-switching frequency-synthesizer technology. Advancing these and other technologies is part of the overall mission to find that all-important tactical edge which inevitably brings an end to an armed conflict.

Jack Browne

Publisher/Editor



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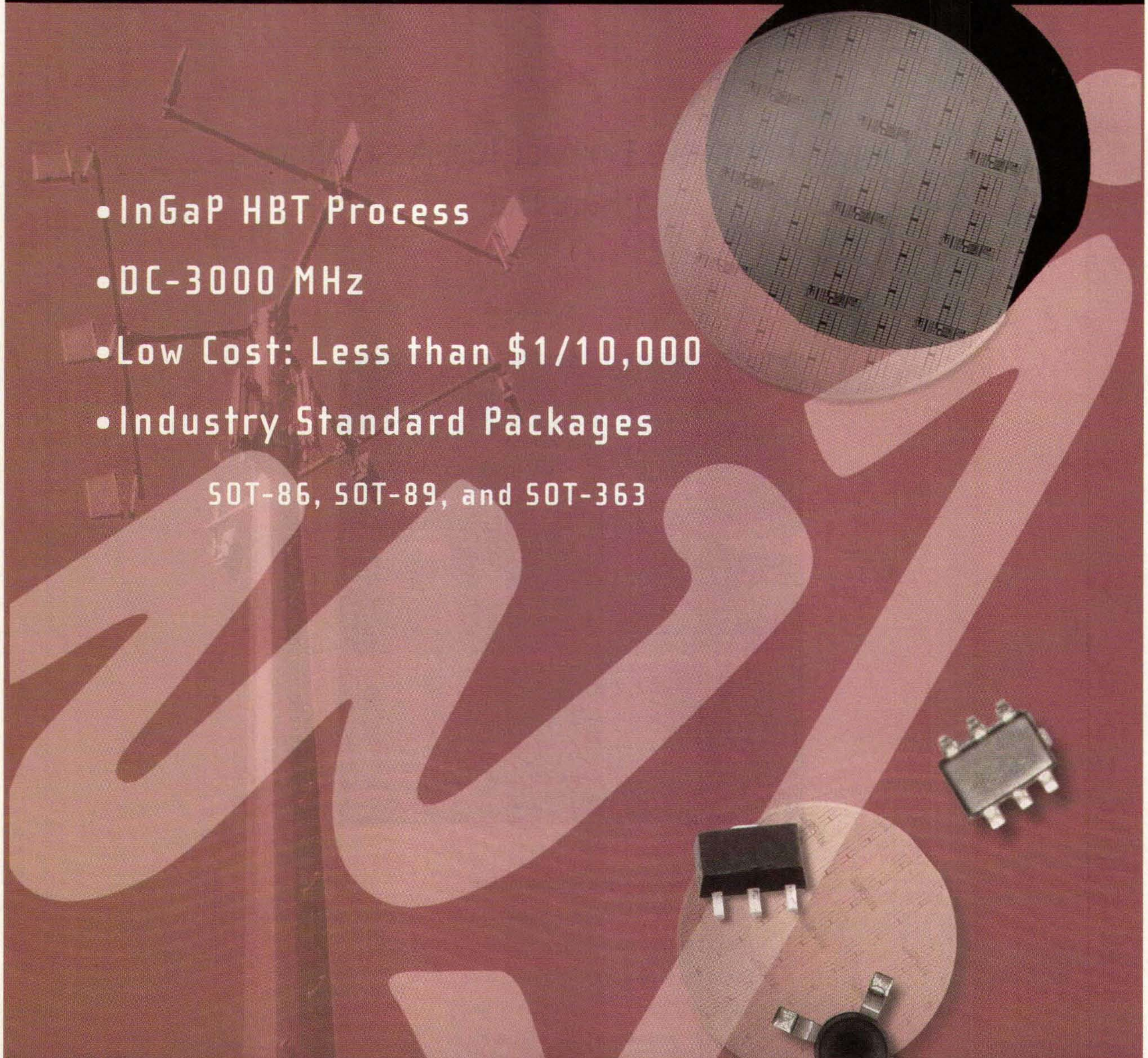

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MODEL NUMBER	GAIN	OUTPUT IP3	P1dB
AG201	11 dB	+20 dBm	+6 dBm
AG202	15 dB	+20 dBm	+7 dBm
AG203	20 dB	+20 dBm	+7 dBm
AG302	15 dB	+27 dBm	+13 dBm
AG303	20.5 dB	+27 dBm	+13 dBm
AG402	15 dB	+33 dBm	+17 dBm
AG403	20.5 dB	+32 dBm	+17 dBm
AG503	19 dB	+29 dBm	+15 dBm
AG602	14 dB	+33.5 dBm	+18.5 dBm
AG603	17.5 dB	+33.5 dBm	+18.5 dBm
AG604	21 dB	+33 dBm	+18 dBm

Specifications listed are typical at 900 MHz

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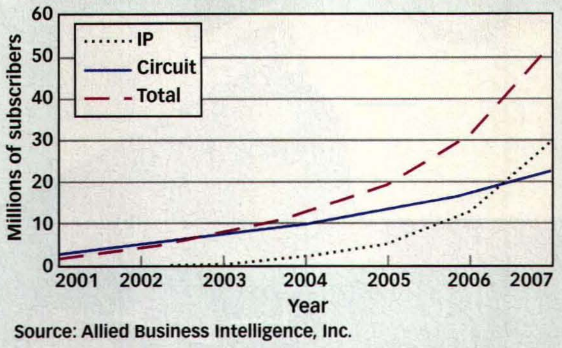
Cable Telephony Is Exceeding Industry Expectations

OYSTER BAY, NY—The growth of cable telephony has exceeded most industry forecasts for 2001. Cable operators from around the world are continuing to deepen their investments in circuit-switched technology while transitioning to Internet-protocol (IP) cable telephony. Though IP reduces cost and improves network efficiencies, equipment vendors have yet to meet the expectations of cable operators. In a report titled "Cable Telephony: The Transition from Circuit Switched To IP—Worldwide Deployments and Opportunities for Equipment Vendors," Allied Business Intelligence (ABI) identifies key requirements that cable operators are looking for before they deploy IP cable-telephony services nationwide. According to ABI, AT&T Comcast, Optus, Cox, UnitedGlobalCom, and Jupiter Communications continue to lead the industry in cable-telephony subscribers for 2002.



ABI's research also indicates: For year ending 2002, there will be 5.21 million cable-telephony subscribers compared to 2.93 million in 2001 **(see figure)**; Revenue generated by cable operators for cable telephony for year ending 2002 will be $2.11 billion dollars; and more mergers and acquisitions to follow the AT&T Comcast merger in the cable industry.

"Cable operators can now offer a bundled package of voice, data, and video, which significantly lowers churn rate, while increasing revenue per subscriber," said John W. Chang, the report's author and senior analyst at ABI.

The First Bluetooth Interoperability Lab In Europe Is Opened

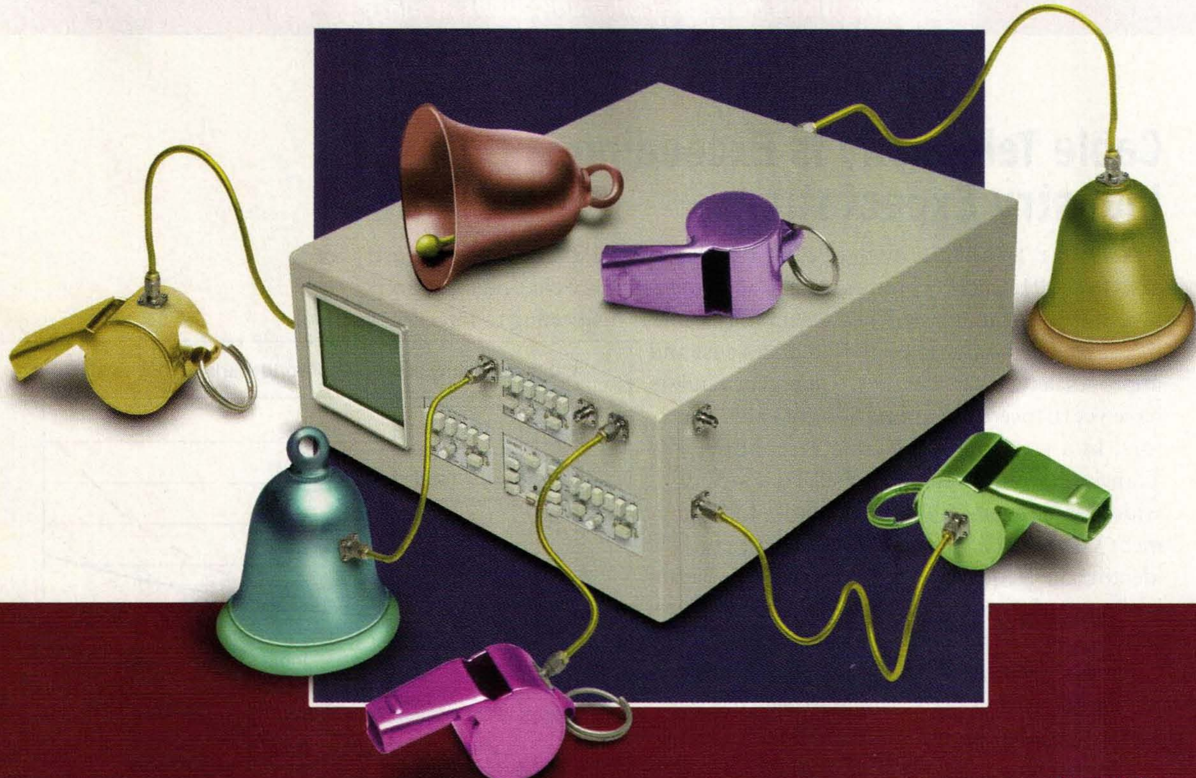
MALAGA, SPAIN—Centro de Tecnologia de las Comunicaciones, S.A. (CETECOM) has announced that the first Bluetooth Interoperability Testing Laboratory in Europe has been set up. FUJITSU Ltd. has joined the program with a Bluetooth-enabled notebook. This *Reference Solution* is the LifeBook B-2547. CETECOM provides FUJITSU Ltd. with testing and consultancy services, to qualify and approve FUJITSU's Bluetooth products for international markets.

The interoperability lab, located at CETECOM, provides a space for companies to test their Bluetooth products' ability to operate across different platforms and with a wide variety of similar devices. The center has been entirely developed by CETECOM to create a "one-stop" research and test facility for all wireless networking interoperability issues. The idea is to have products from many companies that can be tested against the so-called *Reference Solutions* to determine whether or not they are compatible.

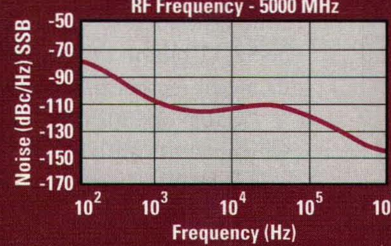
"CETECOM decided to create this laboratory to promote the Bluetooth Wireless Technology, helping developers through interoperability issues in addition to the Qualification Process," explained Rafael Garcia, Bluetooth Manager for CETECOM. "When we receive the Reference Solutions, they are verified by our experts and then interworking with equipment from other companies is analyzed and reported."

"Interoperability testing is the key factor to ensure that one Bluetooth device works properly with another Bluetooth device," continued Garcia.

Do you really need all the Bells and Whistles?



High performance frequency synthesizers give you the performance you want without the extra cost of options you don't need.



Micro Lambda Wireless, Inc. a leader in the development of next-generation YIG devices introduces a new line of high performance frequency synthesizers covering the 600 MHz to 10 GHz frequency range. Designed specifically for wide band and low noise applications, these new frequency synthesizers rival the best lab-grade test instruments on the market.

MLSW-SERIES WIDE BAND FREQUENCY SYNTHESIZERS.

This series of frequency synthesizers offers standard Multi-Octave tuning ranges covering 600 MHz to 3 GHz, 2 GHz to 8 GHz and 2 GHz to 10 GHz. Output power levels of between +10 dBm and +12 dBm are offered depending on frequency band. Frequency step size of 1 Hz is standard, but is programmable with software for customer specific requirements. External reference frequency of 10 MHz is utilized, but 5 to 50 MHz are offered as options. Excellent phase noise performance at 10 kHz offset of -110 dBc/Hz, -108 dBc/Hz and -106 dBc/Hz are provided for the 0.6 GHz to 3 GHz, 2 GHz to 8 GHz and 2 GHz to 10 GHz units respectively. The units operate from +15 Volt and +5 Volt supply lines and frequency control is via a 5-wire serial (SPI & busy) input protocol. Options include dual RF outputs and/or an L-band 2nd L.O. All units measure 5" x 7" x 1" and weigh 28 oz.

FEATURES

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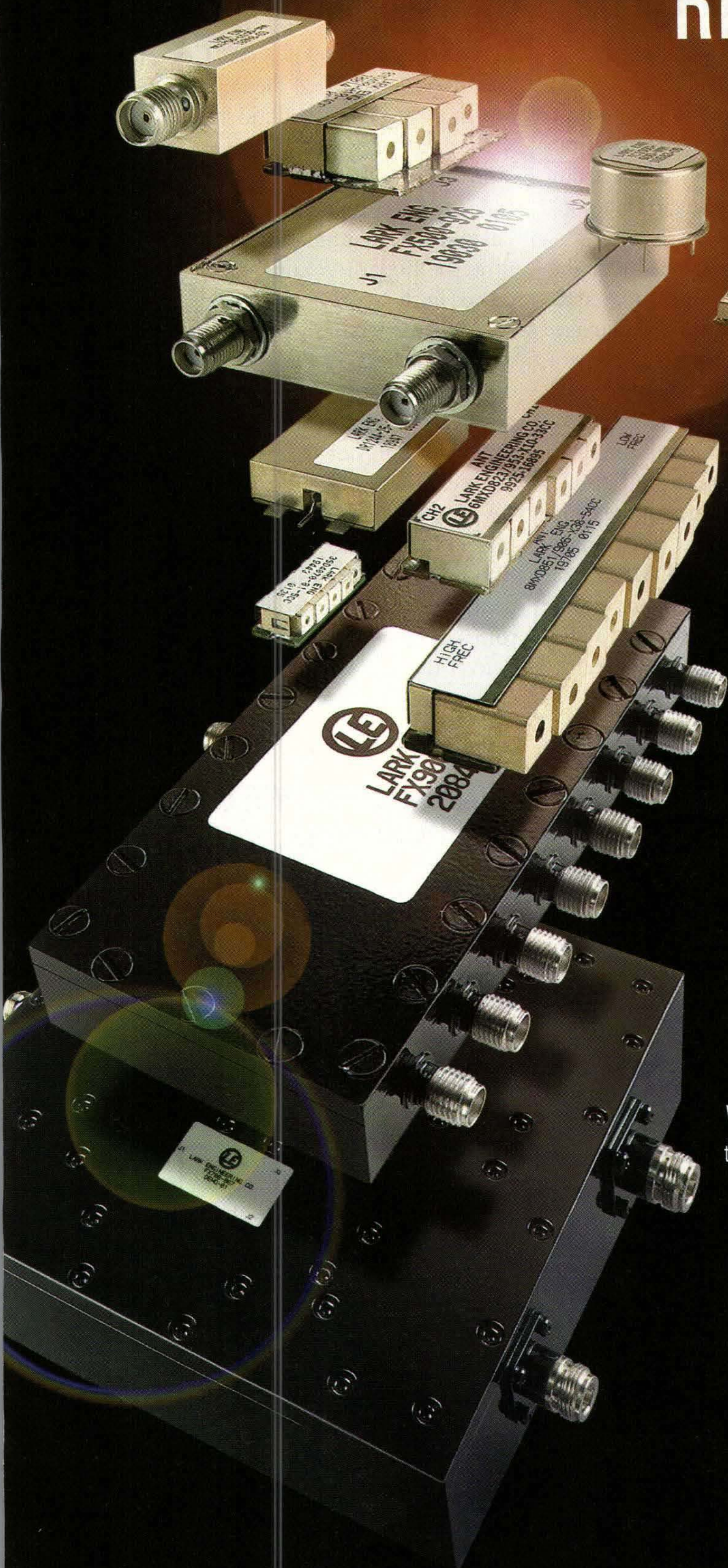

"Communications companies that can leverage CRM will have a significant advantage."

Worldwide Mobile-Phone Sales Declined By 3.8 Percent In The First Quarter Of 2002

SAN JOSE, CA—Worldwide mobile-phone-unit sales reached 93.8 million units in the first quarter of 2002, a 3.8-percent decline from the first quarter of 2001, according to Dataquest, Inc., a unit of Gartner, Inc.

Market-leader Nokia experienced a slight decline in global sales to end-users compared to the same quarter last year. Nokia did manage to register a slight increase in market share versus the same quarter in 2001, despite very weak market conditions in some of its core markets.

Motorola's market-share position continued to grow strongly in the first quarter of 2002, thanks to its continued dominance of the Chinese mobile terminal market and its strength in code-division-multiple-access (CDMA) markets worldwide. Samsung and Siemens experienced the strongest increase in sales, with growth rates of 48.6 percent and 24.1 percent, respectively, during the quarter.

"Samsung's spectacular growth is a reflection of its ongoing success in delivering compelling products across multiple technologies in disparate markets," said Bryan Prohm, senior analyst with the Mobile Communications Worldwide research group for Gartner Dataquest. "Siemens, meanwhile, continued to build on the success it achieved during the latter half of 2001, and it is positioned to make a competitive push in the North American and Latin American GSM markets in 2002."

Mobile terminal consumption in several key regional markets was again disappointing in the first quarter of 2002. In particular, both Western Europe and Latin America suffered their second-consecutive first-quarter fall in year-over-year sales. The Asia/Pacific region (excluding Japan) however, exceeded expectations, as sales in the first quarter increased by 8 percent versus the first quarter of 2001, and jumped by more than 12 percent sequentially from the fourth quarter of 2001.

"Regional demand was highest in Asia/Pacific, where Chinese Lunar New Year promotions and heavy, unofficial terminal subsidies in the Korean market served as catalysts for a strong sell-through," commented Ann Liang, industry analyst with the Mobile Communications Worldwide research group for Gartner Dataquest in Asia/Pacific.

The World's Fastest Computer System Debuts

LOS ANGELES, CA—Michael's Computers and Intel have introduced the Pentium® 4 processor at 2.53 GHz—or 2.53 billion cycles per second.

Michael's Computers' Light-Speed Technology causes no delay with any function calculated. The Pentium 4 2.53-GHz processor is now the highest-volume shipping microprocessor in the world.

"The leadership performance and higher density of the 2.53-GHz Pentium 4 processor are the results of continuous improvements in design and process technology," said Sunlin Chou, senior vice president and general manager of the Technology and Manufacturing Group. "We have enhanced our 0.13-µm process with faster transistors, smaller feature sizes, and 300-mm manufacturing efficiencies, while we continue to ramp up production in multiple factories. This is part of our ongoing effort to deliver the best computing capabilities to our customers in volume."

CRM Initiatives Yield Positive Results For Telecom Firms

EDISON, NJ—Telecommunications companies in North America are realizing tangible business benefits from customer-relationship-management (CRM) initiatives and have a higher level of CRM activity than most other industries. These are just two of the findings of a recent Fujitsu Consulting research study.

The research—which included a survey of 45 communication companies and case studies on eight major telecommunications firms—found that several leading communications had achieved major improvements in the efficiency and effectiveness of their sales, marketing, and service processes through skillful implementation of CRM technology.

"Three forces have been driving communications companies to aggressively pursue CRM: soaring costs of marketing and sales, increasing consumer churn, and rising competition," commented David Yamashita, Telecom 360 Solutions Director at Fujitsu Consulting. "These forces are likely to increase as more and more wire-line traffic moves to wireless and Internet networks. Communications companies that can leverage CRM will have a significant advantage.

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The introduction of new wireless standards has encouraged the development of new linear power amplifier architectures. Vari-L is introducing a line of 3 dB hybrid couplers to facilitate PA engineers' design challenges. Our couplers offer superior phase balance, excellent modeled-to-measured correlation and power handling up to 150W.

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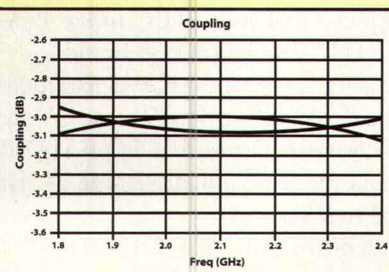
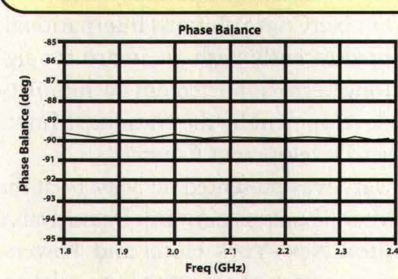
Part Number	Coupling Frequency	Start Frequency	Stop Frequency	Amplitude Balance (dB max.)	Insertion Loss (dB max.)	Phase Balance (Degrees max.)	Isolation (dB min.)	VSWR (max:1)	Power Handling (Watts)	Package
AM03M	3	1700	2000	+/- 0.2	0.20	2	23	1.17	60	0.56"x 0.20"x 0.072"
AP03M	3	2000	2300	+/- 0.2	0.20	2	23	1.17	60	0.56"x 0.20"x 0.072"
AW03M	3	2300	2700	+/- 0.2	0.20	3	22	1.18	60	0.56"x 0.20"x 0.072"
BC03M	3	3300	3700	+/- 0.2	0.20	4	22	1.19	60	0.56"x 0.20"x 0.072"
AH03L	3	815	960	+/- 0.3	0.23	3	22	1.18	150	0.56"x 0.35"x 0.075"
AN03L	3	1500	2200	+/- 0.4	0.25	3	20	1.20	100	0.56"x 0.35"x 0.075"
AR03L	3	1800	2200	+/- 0.2	0.25	3	20	1.20	100	0.56"x 0.35"x 0.075"
AV03L	3	1800	2700	+/- 0.5	0.30	5	18	1.25	60	0.56"x 0.35"x 0.075"
AS03L	3	1930	1990	+/- 0.15	0.23	2	21	1.17	100	0.56"x 0.35"x 0.075"
AP03L	**3**	**2000**	**2300**	**+/- 0.2**	**0.20**	**2**	**23**	**1.17**	**60**	**0.56"x 0.35"x 0.075"**
AY03L	3	3400	3500	+/- 0.3	0.30	5	21	1.25	60	0.56"x 0.35"x 0.075"

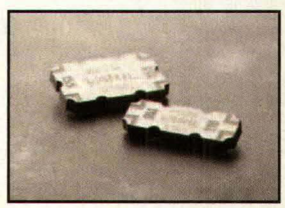
Actual data for AP03L

Coupling



Phase balance





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Gabriel Implements Aggressive Restructuring Plan

SCARBOROUGH, ME—Robert Cameron, the CEO of Gabriel Electronics, Inc., has detailed a restructuring plan that sharpens the firm's focus on market and product strengths that have been at the heart of its success with its licensed and unlicensed wireless broadband antenna solutions for Internet access, mobile, and telephony networks. The restructuring is a direct result of the overall reduction in capital expenditures for infrastructure products by Gabriel's customers worldwide.

"Our top priorities guiding this restructuring are the rapid delivery of Gabriel's innovative antennas to our loyal end-user, OEM, and distribution channels, while remaining focused on our development of the next-generation QuickFire™, FlatFire™, and GemFire™ antenna-solution platforms that carry us into the future," Cameron stated. "We have positioned ourselves to best achieve these goals by streamlining our organization, simplifying our product lines, and targeting investments in activities that are central to our core strengths."

"We have positioned ourselves to best achieve these goals by streamlining our organization."

Gabriel's organization is being streamlined with two goals: first, to continue bringing performance-oriented products to market quickly, delivering them aggressively to customers in Gabriel's key markets; and second, to deliver the company's QuickFire, FlatFire, and GemFire antenna solutions that should launch the company forward.

Kudos

NEW YORK, NY—Emmit J. McHenry, chairman and CEO of NetCom Solutions International, was honored recently with an award for significant long-term achievement in minority business development by the National Minority Supplier Development Council.

The award was presented on May 14 at the annual NMSDC Leadership Awards Dinner-Dance at the Hilton New York Hotel and Towers. The event was part of the year-long celebration of NMSDC's 30th anniversary. More than 1200 people, including CEOs and executives of Fortune 500 corporations and minority business owners from across the nation, attended the event. NetCom Solutions is an integrated logistics and network integration company. It provides networking and supply-chain management solutions to telecommunications, aviation, automotive, and manufacturing organizations nationwide. The company's headquarters are in Chantilly, VA; its national service center is located in Oklahoma City, OK.

In addition to McHenry, awards went to Gerald F. Diez, chairman and CEO of The Diez Group, and Ivan Seidenberg, president and CEO of Verizon Communications. McHenry was recommended by AT&T to NMSDC's Corporate Plus Program.

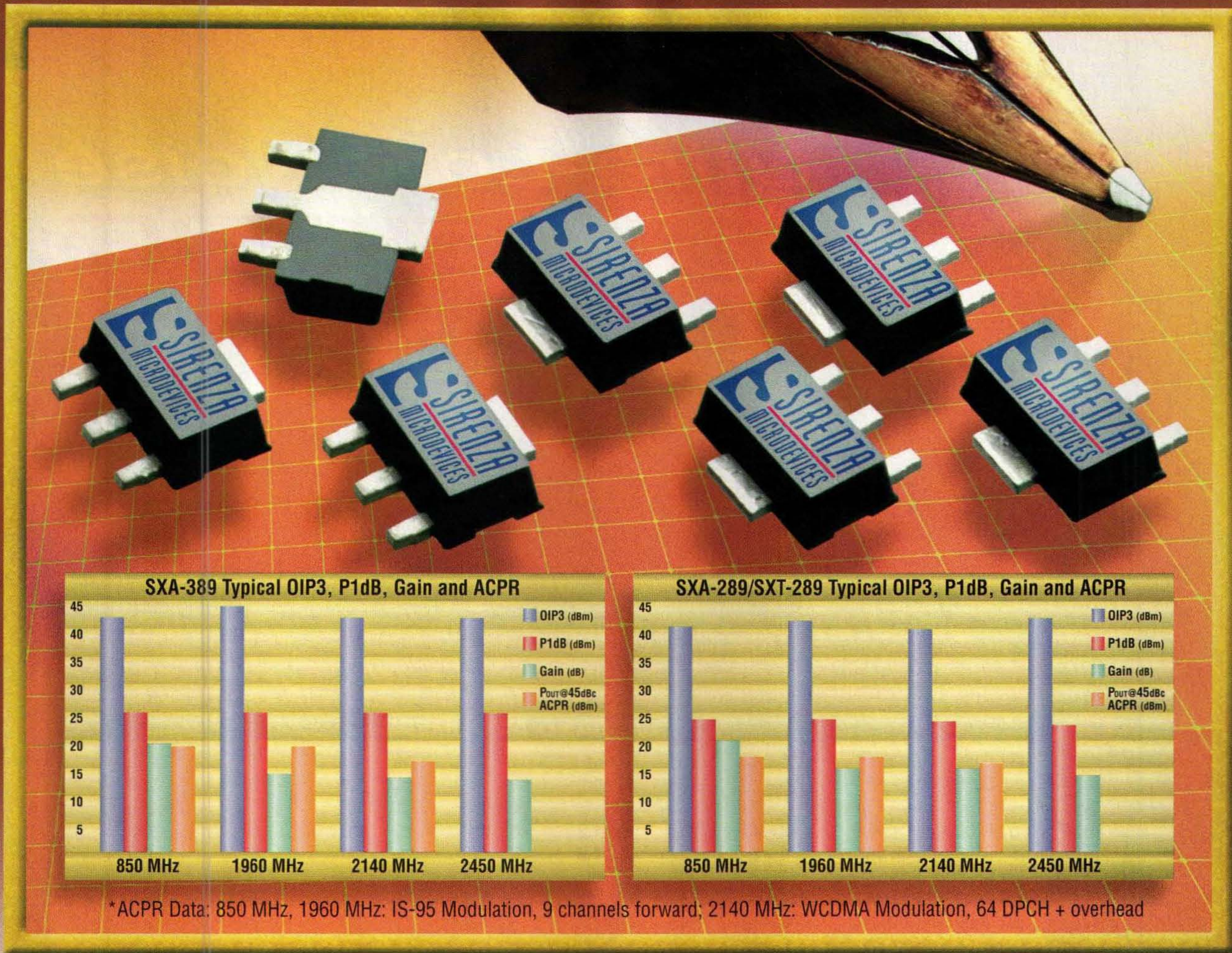
SUNNYVALE, CA—Wireless solutions provider Royal Philips Electronics has unveiled the industry's first 802.11a wireless-local-area-network (WLAN) solution to receive the CE Mark for the European Union market. Based on a rigorous testing process—in consultation with a European Notified Body—Philips now uses the CE Mark, which indicates that the company's WLAN solution meets the requirements of the Radio & Telecommunications Terminal Equipment Directive and will be sold throughout most of the European Union.

The CE Marking of this 802.11a device indicates that it meets European radio-conformance requirements, including dynamic-frequency-selection (DFS) and transmit-power-control (TPC) functionality. Philips' 802.11a solution is the first product meeting these requirements to be launched in a majority of countries in Europe.

INDIANAPOLIS, IN—E-A-R Specialty Composites, a developer and manufacturer of noise- and vibration-control materials, announced that the company's urethane manufacturing plant in Newark, DE has received QS 9000 certification, meeting the stringent quality requirements set by that certification process. The Newark facility produces a range of proprietary energy-control materials under E-A-R's TUFCOTE® and ISOLOSS® brands.

HANOVER PARK, IL—In a cocktail reception held during the IWCE in Las Vegas, NV in late March, Steven L. Deppe, the CEO of MAXRAD, awarded the fifth annual George M. Hanus Award to William C. Mueller. Established in 1998 in memory of MAXRAD's founder, the award seeks to recognize a leadership figure within the wireless communications industry. Mueller served as president and CEO of Hutton Communications for 20 years.

As part of the award, MAXRAD provides a $1000 scholarship to a student at a technical college or university of the recipient's choice. MRF



Linearity, Power... and Price

A powerful HBT amplifier combination.

Sirenza Microdevices' highly linear GaAs HBT amplifiers deliver ¼-watt power with a third-order intercept point as high as 44 dBm. And the prices make them the best buy around.

The SXA-389 runs on 5 volts and offers on-chip active bias control and excellent DC power efficiency. It offers IS-95 channel power of 19 dBm and WCDMA channel power of 16.5 dBm at -45 dBc adjacent channel power. Designed specifically as a driver for infrastructure equipment and customer-premise equipment in the 400–2500 MHz cellular, ISM, WLL, PCS and WCDMA bands, it's priced at just $4 each in quantities of 10,000.

SOT-89 Package shown *actual size!*

The SXA 289 and SXT-289 amplifiers cover the 5–2000 MHz and 1800–2500 MHz bands with a rare combination of efficient ¼-watt power with high linearity in a low-cost, surface mountable SOT-89 package. Both products feature SMDI's high-reliability HBT technology and deliver high OIP3 performance of better than 40 dBm. The price in quantities of 10,000 is just $3.50 each.

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Ultra-Low Noise

AMPLIFIERS

VHF To V-BAND

MODEL NUMBER	FREQUENCY RANGE (GHz)	GAIN (dB, Min.)	GAIN VARIATION (±dB, Max.)	NOISE FIGURE (dB, Max.)	VSWR IN	VSWR OUT	POWER OUT @ 1 dB COMP. (dBm, Min.)	DC POWER @ +15 V (mA, Nom.)
OCTAVE BAND AMPLIFIERS								
JS2-00500100-045-5A	0.5 – 1	35	1	0.45	2:1	2:1	5	250
JS2-00500100-12-5A	0.5 – 1	35	1.2	1	2:1	2:1	5	250
JS2-01000200-045-5A	1 – 2	33	1	0.45	2:1	2:1	5	250
JS2-02000400-045-5A	2 – 4	28	1.2	0.45	2:1	2:1	5	175
JS2-04000800-08-0A	4 – 8	22	1.2	0.8	2:1	2:1	0	150
JS3-04000800-08-5A	4 – 8	30	1	0.8	2:1	2:1	5	175
JS3-04000800-15-5A	4 – 8	30	1	1.5	2:1	2:1	5	175
JS2-08001200-11-5A	8 – 12	15	1	1.1	2:1	2:1	5	150
JS3-08001200-11-5A	8 – 12	25	1	1.1	2:1	2:1	5	175
JS3-08001200-15-5A	8 – 12	25	1	1.5	2:1	2:1	5	175
JS3-12001800-16-5A	12 – 18	23	1	1.6	2:1	2:1	5	175
JS4-12001800-145-5A	12 – 18	30	1	1.45	2:1	2:1	5	200
JS4-12001800-30-5A	12 – 18	30	1	3	2:1	2:1	5	200
JS2-18002600-20-5A	18 – 26	14	2	2	2.5:1	2.5:1	5	100
JS2-18002600-30-5A	18 – 26	14	2	3	2.5:1	2.5:1	5	100
JS3-18002600-20-5A	18 – 26	22	1.8	2	2.5:1	2.5:1	5	175
JS3-18002600-30-5A	18 – 26	22	1.8	3	2.5:1	2.5:1	5	175
JS4-18002600-19-5A	18 – 26	33	1.5	1.9	2:1	2:1	5	200
JS4-18002600-26-5A	18 – 26	33	1.5	2.6	2:1	2:1	5	200
JS2-26004000-35-5A	26 – 40	10	2	3.5	2.5:1	2.5:1	5	100
JS2-26004000-45-5A	26 – 40	10	2	4.5	2.5:1	2.5:1	5	100
JS3-26004000-35-5A	26 – 40	18	2.5	3.5	2.5:1	2.5:1	5	175
JS3-26004000-45-5A	26 – 40	18	2.5	4.5	2.5:1	2.5:1	5	175
JS4-26004000-40-5A	26 – 40	23	2.5	4	2:1	2:1	5	200
JS4-40006000-65-0A	40 – 60	15	3	6.5	2.75:1	2.75:1	0	175
MULTIOCTAVE BAND AMPLIFIERS								
JS2-00500200-07-5A	0.5 – 2	32	1	0.7	2:1	2:1	5	295
JS2-00500200-15-5A	0.5 – 2	32	1	1.5	2:1	2:1	5	295
JS2-01000400-08-5A	1 – 4	27	1	0.8	2:1	2:1	5	200
JS2-01000400-20-5A	1 – 4	27	1	2	2:1	2:1	5	200
JS2-02000600-08-5A	2 – 6	22	1	0.8	2:1	2:1	5	125
JS2-02000600-20-5A	2 – 6	22	1	2	2:1	2:1	5	125
JS2-02000800-08-0A	2 – 8	22	1.25	0.8	2:1	2:1	0	125
JS2-02000800-20-0A	2 – 8	18	1.25	2	2:1	2:1	0	125
JS3-02001800-25-5A	2 – 18	23	1.8	2.5	2.5:1	2.5:1	5	150
JS3-02001800-50-5A	2 – 18	23	1.8	5	2.5:1	2.5:1	5	150
JS4-02001800-22-5A	2 – 18	30	2	2.2	2.5:1	2.5:1	5	200
JS4-02001800-50-5A	2 – 18	30	2	5	2.5:1	2.5:1	5	200
JS3-02002600-33-5A	2 – 26	21	2.5	3.3	2.5:1	2.5:1	5	150
JS3-02002600-40-5A	2 – 26	21	2.5	4	2.5:1	2.5:1	5	150
JS3-06001800-16-5A	6 – 18	23	1.8	1.6	2:1	2:1	5	125
JS3-06001800-30-5A	6 – 18	23	1.8	3	2:1	2:1	5	125
JS4-06001800-145-5A	6 – 18	31	2	1.45	2:1	2:1	5	200
JS4-06001800-30-5A	6 – 18	31	2	3	2:1	2:1	5	200

MITEQ's JS SERIES AMPLIFIERS

- High Performance/Price Ratio
- Superior, Rugged Technology
- Low Phase Distortion Design

Actual
18 to 40 GHz Design

MODEL NUMBER	FREQUENCY RANGE (GHz)	GAIN (dB, Min.)	GAIN VARIATION (±dB, Max.)	NOISE FIGURE (dB, Max.)	VSWR IN	VSWR OUT	POWER OUT @ 1 dB COMP. (dBm, Min.)	DC POWER @ +15 V (mA, Nom.)
MULTIOCTAVE BAND AMPLIFIERS (continued)								
JS3-08001800-16-5A	8 – 18	24	1.5	1.6	2:1	2:1	5	150
JS3-08001800-30-5A	8 – 18	24	1.5	3	2:1	2:1	5	150
JS4-08001800-145-5A	8 – 18	32	2	1.45	2:1	2:1	5	200
JS4-08001800-30-5A	8 – 18	32	2	3	2:1	2:1	5	200
JS3-12002600-25-5A	12 – 26	22	2.5	2.5	2.2:1	2.2:1	5	150
JS3-12002600-35-5A	12 – 26	22	2.5	3.5	2.2:1	2.2:1	5	150
JS4-12002600-22-5A	12 – 26	32	2.2	2.2	2:1	2:1	5	200
JS4-12002600-35-5A	12 – 26	32	2.2	3.5	2:1	2:1	5	200
JS3-18004000-38-5A	18 – 40	16	2.5	3.8	2.5:1	2.5:1	5	150
JS3-18004000-50-5A	18 – 40	16	2.5	5	2.5:1	2.5:1	5	150
JS4-18004000-30-5A	18 – 40	23	2.5	3	2.5:1	2.5:1	5	200
JS4-18004000-50-5A	18 – 40	23	2.5	5	2.5:1	2.5:1	5	200
ULTRAWIDE BAND AMPLIFIERS								
JS2-00100200-07-5A	0.1 – 2	32	1	0.7	2:1	2:1	5	295
JS2-00100200-15-5A	0.1 – 2	32	1	1.5	2:1	2:1	5	295
JS2-00100400-08-5A	0.1 – 4	27	1	0.8	2:1	2:1	5	200
JS2-00100400-12-5A	0.1 – 4	27	1	1.2	2:1	2:1	5	200
JS2-00100600-10-3A	0.1 – 6	23	1.5	1	2:1	2:1	3	175
JS2-00100600-20-3A	0.1 – 6	23	1.5	2	2:1	2:1	3	175
JS2-00100800-13-0A	0.1 – 8	20	1.5	1.3	2:1	2:1	0	175
JS2-00100800-25-0A	0.1 – 8	20	1.5	2.5	2:1	2:1	0	175
JS3-00101000-20-5A	0.1 – 10	23	1.5	2.0	2.5:1	2:1	5	150
JS3-00101000-35-5A	0.1 – 10	23	1.5	3.5	2.5:1	2:1	5	150
JS3-00101200-21-5A	0.1 – 12	23	1.5	2.1	2.5:1	2:1	5	150
JS3-00101200-35-5A	0.1 – 12	23	1.5	3.5	2.5:1	2:1	5	150
JS3-00101800-24-5A	0.1 – 18	23	1.8	2.4	2.5:1	2.2:1	5	150
JS3-00101800-40-5A	0.1 – 18	23	1.8	4	2.5:1	2.2:1	5	150
JS4-00101800-23-5A	0.1 – 18	29	1.8	2.3	2.5:1	2.2:1	5	200
JS4-00101800-40-5A	0.1 – 18	29	1.8	4	2.5:1	2.2:1	5	200
JS4-00102000-25-5A	0.1 – 20	28	1.8	2.5	2.5:1	2.5:1	5	200
JS4-00102000-35-5A	0.1 – 20	28	1.8	3.5	2.5:1	2.5:1	5	200
JS3-00102600-33-5A	0.1 – 26	20	2.5	3.3	2.5:1	2.5:1	5	150
JS3-00102600-42-5A	0.1 – 26	20	2.5	4.2	2.5:1	2.5:1	5	150
JS4-00102600-28-5A	0.1 – 26	27	2.5	2.8	2.5:1	2.5:1	5	200
JS4-00102600-50-5A	0.1 – 26	27	2.5	5	2.5:1	2.5:1	5	200
JS4-00104000-65-5A	0.1 – 40	14	4.5	6.5	2.75:1	2.75:1	5	200
JS4-00104000-85-5A	0.1 – 40	14	4.5	8.5	2.75:1	2.75:1	5	200

For additional information or technical support, please contact either Rosalie DeSousa at (631) 439-9458, e-mail rdesousa@miteq.com or Rizwan Syed at (631) 439-9267, e-mail rsyed@miteq.com

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Generators Deliver Vector Modulation

By combining wideband I/Q modulators and advanced DSP techniques, these instruments offer the power and flexibility needed to produce complex communications waveforms.

Vector signals, with their in-phase (I) and quadrature (Q) signal components, are used to emulate the complex modulation formats found in modern communications systems. Although test-signal generators have been a part of the high-frequency industry from the days of vacuum tubes, vector-signal generators are relatively new tools that are available from only a select handful of suppliers. The key to selecting a vector-signal generator lies in understanding the pertinent specifications and their meaning for specific applications.

JACK BROWNE
Publisher/Editor

The CS2010VSG vector-signal generator is a cross between a PC and a traditional high-frequency signal generator. [Photo courtesy of L3 Communications, Celerity Digital Broadband Test (Cupertino, CA).]

Modern communications systems employ complex digitally modulated signals based on I/Q or vector modulation. In modulation formats such as binary phase-shift keying (BPSK), quadrature phase-shift keying (QPSK), and quadrature-amplitude modulation (QAM), different phase and amplitude states are used to represent digital bits, supporting the transmission of bandwidth-efficient modulation over relatively narrow channel bandwidths.

Many commercial vector-signal generators include a sophisticated blend of hardware and software. The CS2010VSG vector-signal generator from L3 Communications, Celerity Systems Digital Broadband Test (Cupertino, CA), for example, is almost a cross between a personal computer (PC) and a traditional high-frequency signal generator **(see figure)**. It provides operators with a 1-dB channel bandwidth of 30 MHz that can be tuned across a total frequency range (in bands) of 685 to 2200 MHz. Initially designed to generate second-generation (2G) cellular signals, the vector-signal generator now includes a large signal set supporting intermediate-generation (2.5G) and third-generation (3G) testing, including Enhanced Data rates for Global Evolution (EDGE) cdma2000 1X and wideband-code-division-multiple-access (WCDMA) signals. It is also possible to add additive white Gaussian noise (AWGN) to the entire 30-MHz band or only specific portions of the 30-MHz band.

The instrument features direct-to-intermediate-frequency (IF) conversion to produce the desired modulation in software, rather than using the control of higher-frequency I and Q signal components to create the digital modulation. Due to inherent imperfections in the hardware I/Q modulators and baseband I and Q pulse-shaping filters, instruments based on traditional I and Q modulation have a tendency to suffer I versus Q timing errors, I versus Q relative-gain errors, and I versus Q offset errors. The software implementation of the CS2010VSG supports precise control

over I and Q generation.

The CS2010VSG can generate up to 256 carriers, and can be fitted with a variety of output filters, including a Global System for Mobile Communications (GSM) filter for 880 to 960 MHz and a digital-communications-services (DCS) 1800 filter for 1805 to 1880 MHz. The instrument, which is based on a Pentium processor and Windows NT. It includes a file import facility to use existing or newly created I and Q files. These files are converted into direct-to-IF-based signals within the vector signal. Deep arbitrary-waveform-generation (AWG) memory provides up to 23 s (4 GB or 2 GSamples) of a 30-MHz-wide modulation bandwidth.

Along the lines of more traditional "rack-and-stack" test instruments, the SMIQ series of vector-signal generators from Rohde & Schwarz (Munich, Germany) [and available in North America from Tektronix, Inc. (Beaverton, OR, www.tektronix.com)] provide frequency coverage through 6.6 GHz. The vector-signal-generator line currently includes four models with bandwidths accommodating most wireless applications: SMIQ02B (300 kHz to 2.2 GHz), SMIQ03B (300 kHz to 3.3 GHz), SMIQ04B (300 kHz to 4.4 GHz), and SMIQ06B (300 kHz to 6.6 GHz). The SMIQ series instruments share a high-performance I/Q modulator and powerful digital-signal-processing (DSP) technology to generate analog and digital modulation formats. The instruments can generate amplitude modulation (AM) and frequency modulation (FM) as well as an assortment of digital modulation formats, from amplitude shift keying (ASK) to 256-state QAM. Digital filters enable the definition of approximately any type of baseband filtering, and symbol rates to 18 Msymbols/s are available.

The SMIQ vector-signal generators offer a fast settling time of less than 3 ms for frequency and less than 2.5 ms for amplitude. The generators support frequency hopping at rates of 500 μs, and provide users with control of frequency and level sweeps. Signal spectral purity is outstanding, with single-sideband (SSB) phase noise of –126 dBc/Hz offset 20 kHz from a 1-GHz carrier. The level accuracy is ±0.5 dB for output levels to +13 dBm (to +16 dBm in over-range mode). The generators are supported by a wide range of options, including an option for a noise generator and distortion simulator with noise bandwidths that are selectable from 10 kHz to 10 MHz, an option for fading simulation capability (six-path simulation or two generators can be linked for form a 12-path simulation) with selectable path attenuation and delay characteristics and calibrated RF levels from –140 to –5 dBm.

In addition to the vector-signal generators, the company also offers the AMIQ series of I/Q modulation gen-

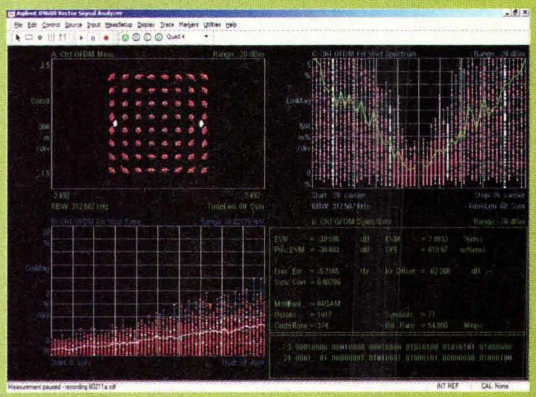
The original idea was simple: use wireless links to replace the tangle of cables that connect PCs, PDAs, mobile phones and more. Of course, turning that idea into reality—without much time for analysis—has been anything but simple. Perhaps we can help.

As you work to create the *Bluetooth* and Wi-Fi 5 products of today and tomorrow, we're busy working with standards committees and engineers like you to address challenges such as interoperability, certification and quality. Here are three quick samples of the insights we've gathered along the way.

Ensuring interoperability.

Many people attribute the popularity of Wi-Fi devices to WECA testing that certifies the interoperability of products from multiple vendors. Of course, the roots of interoperability reach back to the early stages of product development when each manufacturer (or silicon supplier) adds value by optimizing its designs in unique ways. It's a good practice, but one that may leave your products working well together but not with devices from another maker.

Developers tell us interoperability is often a matter of tweaking a transmitter or receiver design. For transmitters, error vector magnitude (EVM) versus time or channel is a measure of modulation quality that can highlight underlying problems such as nonlinear distortion, phase noise and spurious signals. Conversely, one way to make receivers more forgiving of nonideal transmitters is to test them with a variety of impaired signals—in hardware with a flexible signal-generation solution, in a computer-based simulation, or in a system that links both methods.



For this IEEE 802.11a signal, the overall EVM measurement is acceptable but viewing EVM versus time (lower left) and channel (upper right) shows the effect of a timing error.

Clearing the qualification hurdle.

Getting through certification and regulatory testing quickly is obviously a major challenge. With WLAN, it's often useful to focus pre-qualification work on the "PHY" (RF) layer because certification tests it indirectly. Digitized recording simplifies the analysis of problem signals, and replaying captured transmissions from other devices enables repeatable PHY tests.

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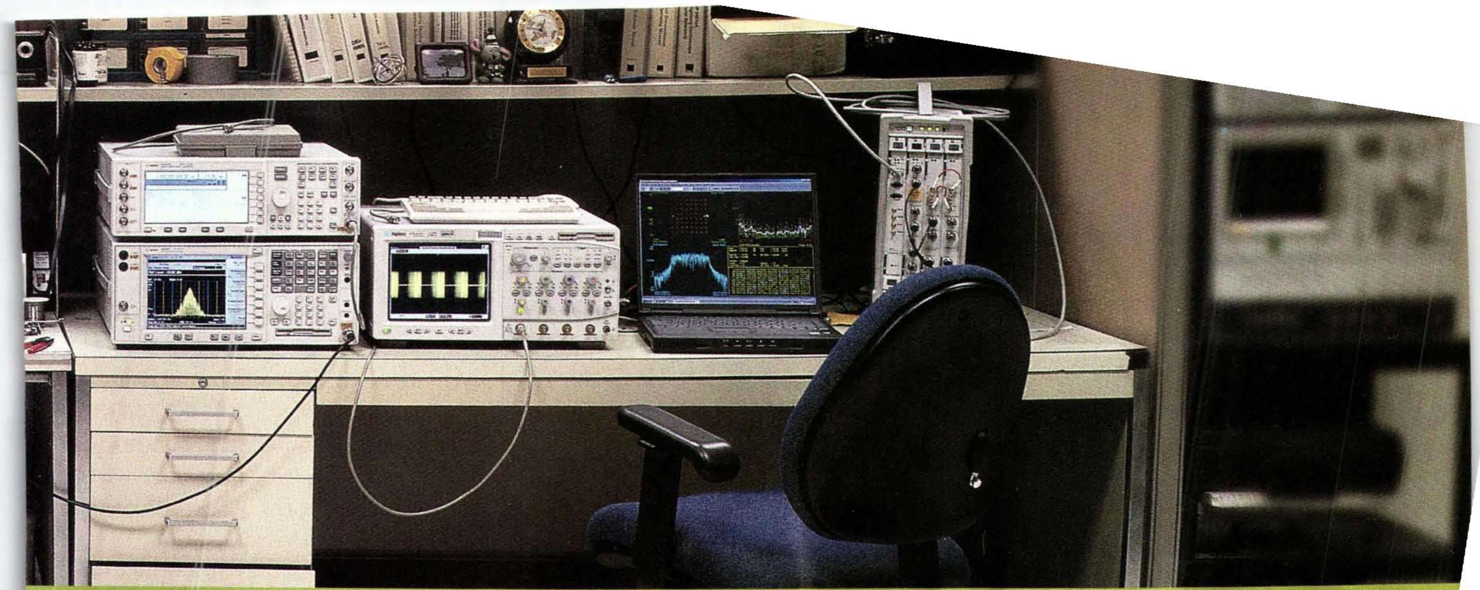

To achieve *Bluetooth* qualification, your product must pass all 16 sections of the RF test specification. From what we've seen, test results are as unique as every design and no single test yields consistent failures. However, most receivers exceed the sensitivity specification so the nine transmitter tests tend to be the biggest concern.

Problems are often specific to the type of transmitter. In FM-based designs, frequency drift is a common issue. Digital noise on the power supply can affect modulators and VCOs, producing ripple in a frequency drift measurement. For IQ-based modulators, the culprit is often modulation quality. They're not called out in the test specification, but measurements such as FSK error, magnitude error and the eye diagram help identify modulation quality problems.

The FSK error display can highlight the effects of unwanted frequency modulation, which may indicate the presence of spurious signals in the modulator.

Optimizing manufacturing test.

Although many Wi-Fi and *Bluetooth* products are designed for consumer applications, most are complex enough to warrant some level of testing on the manufacturing line. But how much testing?

Combining your device expertise with our instrument knowledge, we can create an optimized test program that needs only a subset of the relevant RF test specs. If common test modes are designed in, it's also possible to accelerate some of the tests. For OEMs who purchase and integrat *Bluetooth* subsystems, testing can focus on the PHY layer rather than the protocol layer.

Getting test of the story.

Sharing its and test practices is just one way Agilent can help rate the drive to market with new wireless networkucts. The Agilent Interoperability Certification Labs and's network of test partners are ready to help, to tested hundreds of Wi-Fi devices and can offer vs into clearing the qualification hurdle.

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Anologies

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erators. Offering as much as 16-b resolution at sample rates to 100 MHz, these modulation sources can be used along with an RF generator to create digitally modulated test waveforms.

The E4438C ESG vector-signal generator series from Agilent Technologies (Palo Alto, CA) now offers five frequency ranges, from 250 kHz to 1 GHz, 2 GHz, 3 GHz, 4 GHz, or 6 GHz. The generators feature 160 MB (32 MSamples) baseband memory for waveform playback and 6 GB (1.2 GSamples) of nonvolatile memory for storing waveforms and instrument settings.

The E4438C switches frequency in less than 14 ms and amplitude in less than 19 ms. It delivers +17-dBm output power at 1 GHz with level accuracy of ±0.5 dB. Using external sources with its I and Q input ports, the E4438C supports an RF modulation bandwidth of 160 MHz. With its internal baseband generator, the RF modulation bandwidth is 80 MHz.

The MG3672A digital-modulation signal generator from Anritsu Co. (Morgan Hill, CA) operates at carrier frequencies up to 2.75 GHz. It also features wide-range external I/Q input ports (a 3-dB bandwidth of 30 MHz) for generating a wide range of digitally modulated signals.

A vector-signal-generator solution from Nova Engineering (Cincinnati, OH) uses a programmable-logic-device (PLD) architecture to generate complex modulation by playing back high-resolution digital samples. The generator combines the company's Constellation PLD development board with its numerically controlled oscillator (NCO) intellectual-property (IP) core. Frequency resolution is 32 b, while phase offset can be set to 16-b resolution. The output frequency and resolution are a function of the clock frequency. So, for a 50-MHz clock frequency, the output frequency is approximately 20 MHz and the resolution is $50 \text{ MHz}/2^{32} - 0.012$ Hz.

Finally, for those unwilling to trade in their older signal generators, the model 2029 vector modulator for IFR Systems (Wichita, KS) converts an analog-signal generator into a vector-signal generator with frequency range of 800 to 2510 MHz. The instrument offers a 14-b internal arbitrary-waveform generator and an I/Q modulation bandwidth of 10 MHz that can be used to achieve an effective RF modulation bandwidth of 20 MHz.

These instruments represent a sampling of the vector-signal generators currently on the market. Those interested in a more comprehensive listing are invited to visit the *Microwaves & RF* Product Data Directory website at www.m-rf.com. MRF

Directional Coupler Spans 1 MHz To 1 GHz

THE MODEL GLSL20D102 is a 20-dB surface-mount directional coupler that is designed for RF circuitry, broadband, mobile-radio, and satellite-television technologies. The unit operates from 1 MHz to 1 GHz with a typical insertion loss of less than 0.5 dB from 2 MHz to 1 GHz. Directivity typically exceeds 10 dB from 1 MHz to 1 GHz. Operating temperature range is –40 to +125°C. The couplers is suitable for automatic insertion on surface-mount circuits assembled using reflow or vapor-phase soldering. Terminals are formed on the ceramic base from the ends of the coil windings, eliminating solder joints between the coil and the terminals that could open from the heat of circuit assembly. Soldering-heat resistance is 230°C for 5 s. P&A: $1.20 each; 8 to 10 wks.

Sprague-Goodman Electronics, Inc., 1700 Shames Dr., Westbury, NY 11590; (516) 334-8700, FAX: (516) 334-8771, e-mail: info@spraguegoodman.com, Internet: www.spraguegoodman.com.

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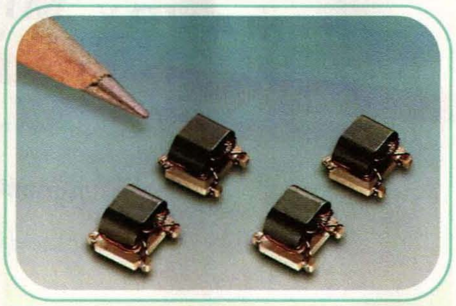

SPRAGUE-GOODMAN MODEL GLSL20D102

Rx Meter Detects Signals From Beyond 80 Ft.

THE ZAP CHECKER is a receiver (Rx) meter with a 10.0-MHz-to-greater-than-4.5-GHz bandwidth that is able to detect signals from cellular phones at more than 20 ft. and ultra-high-frequency (UHF) and very-high-frequency (VHF) transceivers from more than 80 ft. A sensitivity control adjusts the gain beyond the 20-dB range and two antennas are enclosed inside the case. The dynamic range spans a 1000:1 signal landscape in logarithmic mode. Signal readings are from an analog meter or light-emitting-diode (LED) display. A switch-enabled vibration mode is also included. P&A: $89.00.

Alan Broadband Co., Inc., 93 Arch St., Redwood City, CA 94062; (888) 369-9627, Internet: www.zapchecker.com.

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ALAN BROADBAND ZAP CHECKER

Guided-Wave Radar Tx Provides Level Detection

THE MT2000 IS a guided-wave radar-level transmitter (Tx) that measures the level of bulk solids and liquids with an accuracy of ±5.1 mm. The Tx features a metallic probe to act as a wave guide, directing the microwave pulses and eliminating the beam divergence common to conventional, noncontact radar Txs. With no beam divergence, false signal/echoes from objects other than the product surface are eliminated. The probe can be cut to size, from 0.6 to 30.5 m. The unit's output is through 4 to 20 mA, HART, or Honeywell DE. Results are locally indicated with a built-in scrolling liquid-crystal display (LCD) that offers measurements in a field-selected choice of feet, inches, millimeters, centimeters, meters, and percent. A linearization table is included.

K-TEK, 18321 Swamp Rd., Prairieville, LA 70769; (866) 735-5583, FAX: (225) 673-2525, e-mail: sales@ktekcorp.com, Internet: www.ktekonline.com.

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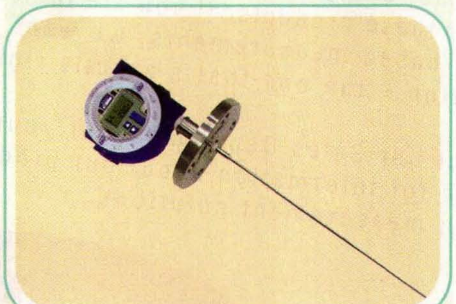

K-TEK MODEL MT2000

VCXO Targets WDM Applications

THE 7311E SERIES of voltage-controlled crystal oscillators (VCXOs) for wavelength-division-multiplexing (WDM) applications incorporates a 155-MHz fundamental crystal, bipolar oscillator circuit, and a selected multiplier/filter circuit. An analog low-phase-noise circuit makes this VCXO a pure-frequency source. The 7311E exhibits frequency stability of ±50 PPM maximum over an operating temperature range of –10 to +70°C. The VCXO operates on a supply voltage of only +3.3 VDC, while consuming just 85 mA. It is available in a frequency range of 622 to 780 MHz. P&A: $34.00 each (10,000 qty. tape and reel); 8 to 10 weeks.

NDK America, Inc., 47671 Westinghouse, Dr., Fremont, CA 94539; (800) 635-9825, (510) 623-6500, FAX: (510) 623-6590, e-mail: sales@ndkxtal.com, Internet: www.ndk.com.

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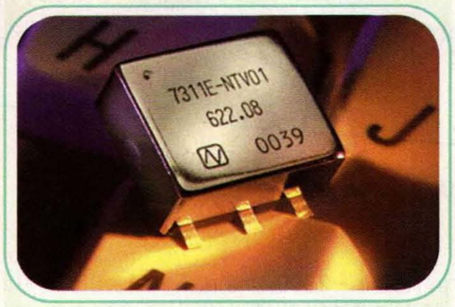


NDK AMERICA, INC. 7311E SERIES

Pager Glory Days Are Over

BEFORE THE cell-phone boom of the mid-to-late 1990s, pagers were one of the main wireless devices for communication. Used mostly by doctors and drug dealers in the early 1990s, the one-way pager is slowly becoming obsolete in the US.

The beeper's glory days were in 1998, when there were 45 million pager users. Since then, that number has dropped by 20 percent. In 2000 (the most recent year for available figures), the number of pager users stood at 37 million.

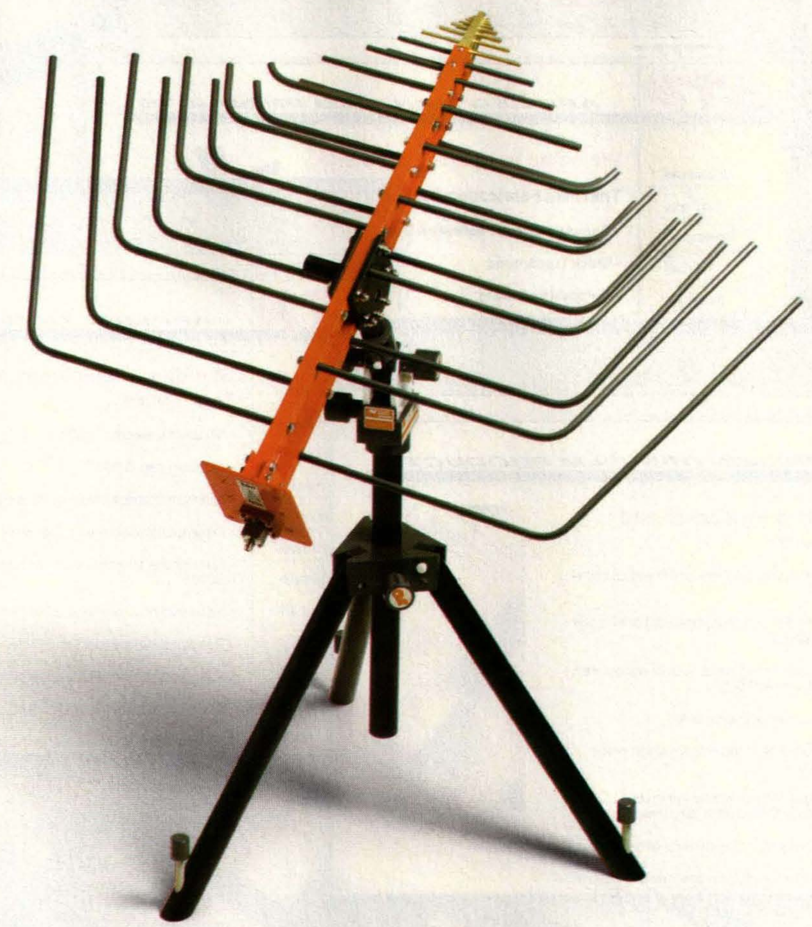
These days, few customers are signing up for pager service. And among those who still have a pager, 25 percent use them little, if at all. In addition, the dominant beeper manufacturer, Motorola, announced last December that it would no longer manufacture the one-way devices.

Although pagers are slowly disappearing, there is a segment of users out there who still use them frequently. Users say that compared with cellular phones, beepers are inconspicuous and allow people to selectively talk to callers rather than being forced to pick up a ringing phone. There is also the cost issue. Pagers are more reliable and less expensive than wireless phones.The average monthly paging bill is approximately $12, while the average cell-phone bill is $61. One user, after hearing the Motorola announcement last December, wrote that he would keep his pager "until they pry it from my cold dead body."

The lower frequency that the pagers use equates to fewer gaps in coverage, because page signals travel farther than signals from cell towers. Hospitals are one of the biggest users of pagers since the lower RF enables the penetration of buildings more easily and the pagers do not interfere with certain pieces of medical equipment, unlike mobile phones.

Beepers started to become obsolete when cell-phone carriers introduced pager-like features such as voice mail and caller ID. Michele Dynan, who has sold pagers in Alexandria, VA since 1990, stated in *The New York Times* that she has lost more than three-fourths of her 7000 customers in the last three years and is looking to shut down her business. "What I'm basically doing now is fixing and replacing pagers," she told the *Times*. MRF

Radiant Arrow



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For more lower-end frequency response, try the other warrior, Radiant Arrow 26™. From 26 MHz to 5.0 GHz.

Both 100% proprietary. Others pale in comparison. Find out more on the web at ***www.amplifiers.com/antbrdbd.asp***

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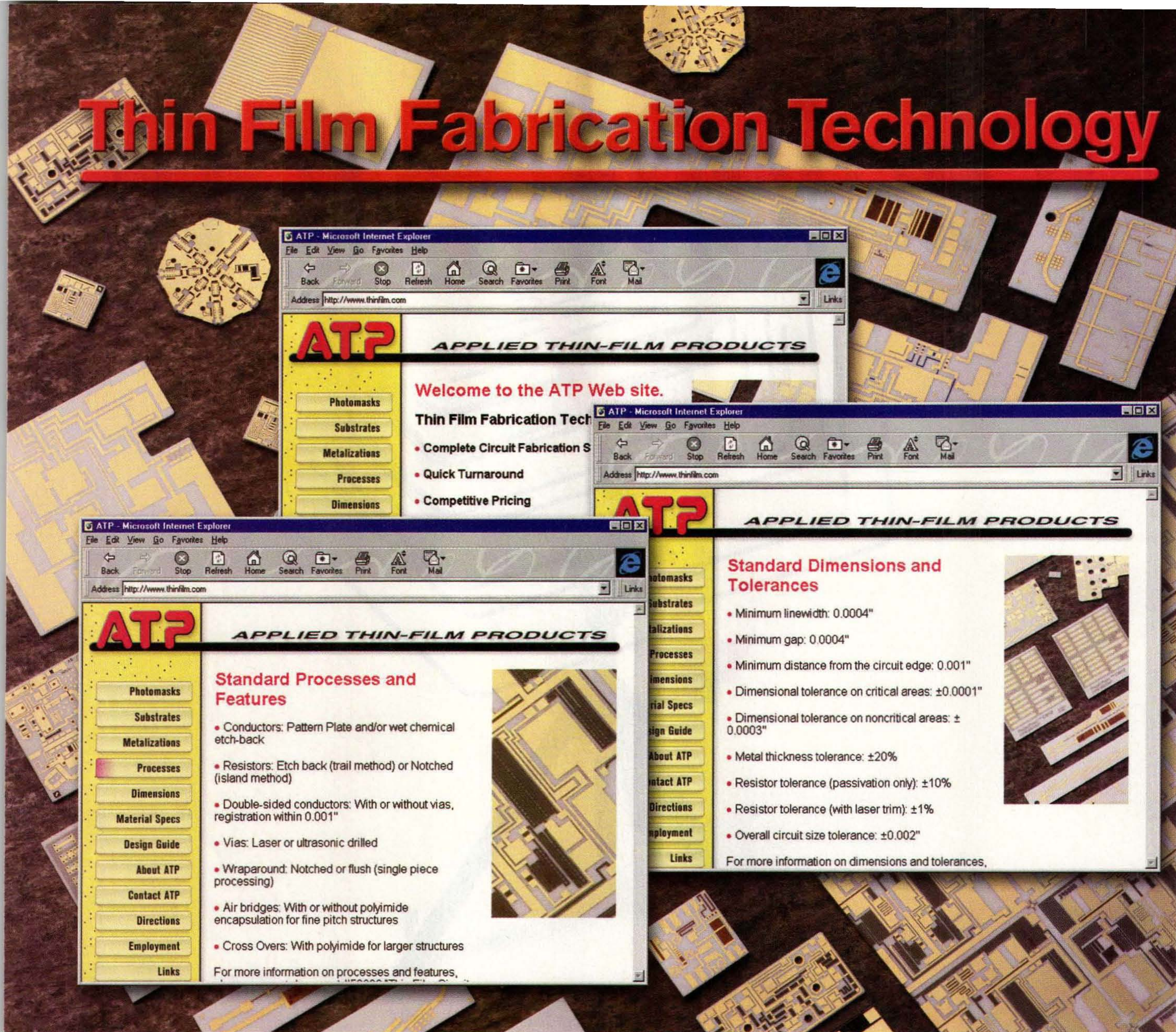

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CONTRACTS

Harris Corp.—Has been awarded a $17 million contract by the US Special Operations Command (USSOCOM). Under the terms of the contract, Harris will supply AN/PRC 150 (C) high-frequency radios from its Falcon® II family of tactical radios.

EMS Technologies, Inc.—Acquired Ottercom Ltd., a provider of Inmarsat communication terminals located in Tewkesbury, UK. The acquisition follows more than 10 years of close collaboration between EMS and the Ottercom organization. The company will now be known as EMS SATCOM UK and will operate as a unit of EMS's SATCOM Division in Ottawa, Ontario, Canada. Financial details of the transaction were not released, although EMS issued 81,245 new shares of its common stock as a portion of the consideration for the transaction.

Andrew Corp.—Announced that it has acquired Quasar Microwave Technology Ltd., a microwave-component company based in Newton Abbot, Devon, England. The addition of Quasar's microwave and millimeter-wave components completes the Andrew RF path offering in terrestrial antenna subsystems for the fixed-line telecommunications market. Financial details of the transaction were not disclosed.

Stratos Lightwave, Inc.—Has entered into a definitive agreement to acquire Paracer, Inc., a privately held company based in Santa Clara, CA that designs and manufactures Parallel Optics transceivers and subsystems. Under terms of the agreement, Paracer will become a wholly owned subsidiary of Stratos.

Electrocube, Inc.—Acquired the product lines of Seacor, Inc. The acquisition of Seacor's products will enable Electrocube to expand their product capabilities as a supplier of an extensive range of capacitor products and custom engineering services for the electronic, communication, industrial, military, and aviation marketplaces.

FRESH STARTS

Paratek Microwave, Inc.—Has raised $17 million in new financing, bringing the company's total capitalization to $53 million. The capital raised will be used to further enhance the commercialization of Paratek's product portfolio. The investor group was led by Morgenthaler Ventures and Novak Biddle Venture Partners. Investors also included Investor AB and DB Capital Venture Partners, as well as a number of undisclosed financial investors.

Net Com Solutions, Inc.—Established a new minority supplier diversity program to benefit its large corporate customers and minority- and women-owned companies as well. It has targeted upward of 3 percent of its revenues in the coming year to minority subcontractors.

TCSI Corp.—Announced the financial results for the first quarter ending March 31, 2002. First-quarter revenues grew 14 percent to $4.2 million, compared to $3.7 million for the first quarter of 2001. The company also reported a first-quarter pre-tax profit of approximately $371,000, compared to a loss of $2.7 million in the first quarter of 2001, representing a net income per share of $0.01. Expenses in the first quarter of 2002 were $3.9 million, down from $6.7 million in the first quarter of last year. The company's cash and cash equivalents also increased by more than $1 million during the quarter.

Proxim Corp.—Revealed financial results for the first quarter ending March 29, 2002. Revenue for the first quarter was $25.4 million, compared to revenue of $37.6 million for the first quarter of 2001, and up sequentially from $24.2 million for the fourth quarter of 2001.

Centurion Wireless Technologies, Inc.—Announced the broad expansion of its customer-service and technical-support capabilities. Centurion's toll-free call center now serves OEMs, distributors, and other Centurion customers coast to coast, throughout their entire business day. Centurion has added additional hours to its customer service and has increased its customer-service staff by 100 percent. To supplement the additional hours and personnel, Centurion has created a new technical support line to be serviced by engineers who will address technical questions and issues. US customers can now call the toll-free sales line at (800) 228-4563 at times convenient to their locale from 6:00 a.m. to 8:00 p.m. Central Time.

Gabriel—Signed a distribution agreement with Sabre Communications Corp. of Sioux City, IA. This non-exclusive agreement authorizes Sabre Communications Corp. to market and supply Gabriel's wireless antenna solutions throughout the US and Canada. Sabre markets communication-related products through its Sabre Site Solutions™ Catalogs, its call center, and website at www.sabrecom.com.

M-tron Industries, Inc.—Appointed a new manufacturer's representative. Glen White Associates (GWA) will handle M-tron's entire line of frequency-control devices in Georgia, North Carolina, South Carolina, and Tennessee. GWA has offices in Huntsville, AL [(256) 882-6751], Duluth, GA [(770) 418-1500], Raleigh, NC [(919) 848-1931], and Huntersville, NC [(704) 875-3777].

Hittite Microwave Corp.—Announced the appointment of two new sales-representative firms to serve customers in Asia. Planet Technology (Hong Kong) Ltd. will handle sales in mainland China and Hong Kong, while MEDs Technologies Pte. Ltd. will cover Singapore, the Philippines, Malaysia, and Indonesia.

Raltron Electronics Corp.—Started production of miniature crystal and oscillator units in its new manufacturing facility in Shanghai, China. The two-story facility is 80,000 sq. ft. and houses production equipment and systems for the manufacture of crystal and oscillator products. MRF

Semflex Names Selogie To VP and GM Position

SELOGIE

Semflex, Inc. has named CAMERON SELOGIE as the company's new vice president and general manager. He is responsible for growth and profit strategies in all functional areas of the operation. He joined Semflex in 1999 as vice president of quality and operations.

Racal Instruments—CARL HIGGINS to business manager for the company's Broadband/Photonics group; formerly business development manager.

Integral Technologies, Inc.—RAVI MIRCHANDANI to vice president of business development; formerly employed at General Electric's Plastics Division.

Omniyig, Inc.—ARISTOMENIS CAPOGEANNIS to marketing manager; formerly employed at Atmel Corp.

Bisco Industries—DON WAGNER to vice president for product management; formerly director of purchasing.

Santa Barbara Infrared—JAY B. JAMES to director of business development; formerly director of marketing and sales at Raytheon Commercial Infrared (RCI).

Interconnect Devices, Inc. (IDI)—JERRY MEISENHEIMER to vice president and CFO; formerly controller. Also, HOWARD WEINER to senior vice president; formerly senior director of operations.

Certicom Corp.—HERVÉ SÉGUIN to CFO; formerly CFO of DWL, Inc.

Institute of Electrical and Electronics Engineers (IEEE) Antennas and Propagation Society (AP-S)—DR. ZOLTAN CENDES to distinguished lecturer for the 2001-02 term; remains as chairman and chief technology officer of Ansoft Corp.

OnLine Power Supply, Inc. (OPS)—NATHAN HUNTER to electrical design engineer; formerly electrical project engineer at Artesyn Technologies. Also, BRYANT ALSTON to senior engineering technician; formerly test and repair supervisor at Carsan Engineering, Inc.

EMS Technologies, Inc.—RAYMOND LARKIN to director of sales and marketing for the SATCOM Aeronautical Group; formerly director of marketing at Teledyne Controls.

Scott Specialty Gases—JACK DE JONG to general manager of Scott's European operations in Breda, The Netherlands; formerly sales and marketing manager.

Palomar Technologies, Inc.—JEFFREY M. KING to vice president of sales and product support; formerly employed in engineering, sales, and marketing management positions at Palomar.

Trompeter Electronics—JOE ANDERSON to field sales engineer for the Dixie States territory of Georgia, Alabama, and Tennessee; formerly regional sales engineer for Atlanta Cable Sales.

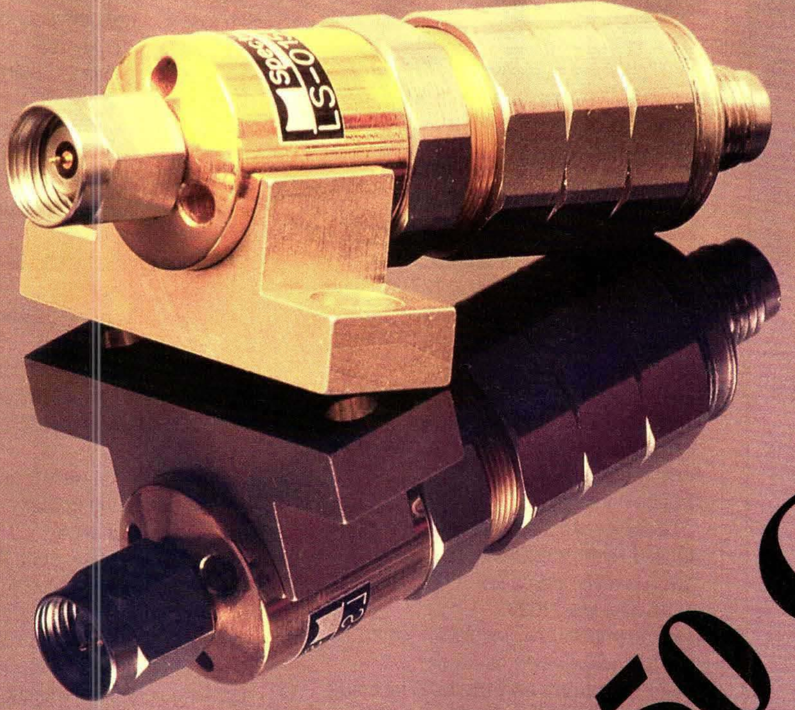
Proxim Corp.—KEVIN J. DUFFY to general manager of the Broadband Services Division; formerly vice president of Home Networking at SIEMENS Information & Communications Mobile. Also, STEVE TIMMERMAN to head of the marketing team; formerly vice president of sales and marketing for Inviso.

TECOM Industries, Inc.—ELISSA SEIDENGLANZ to director of Government Business Development; formerly director of business development for Military Aircraft Electronic Systems at Northrup Grumman's Navigation Systems Division.

SEIDENGLANZ

CLARK

TMD Technologies Ltd.—MIKE CLARK to engineering manager for the Tubes Division; formerly held senior technical and business positions with Marconi. MRF



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Bypass PAAs With A Frequency-Discriminated Phase Controller

AS AN ALTERNATIVE to phased-array antennas (PAAs), Gong-Ru Lin of the Institute of Electro-Optic Engineering, National Taipei University of Technology (Taiwan, R.O.C.) demonstrates the frequency discrimination and phase tuning of high-speed optical clocks generated from a directly modulated/encoded laser diode using a DC-voltage-controlled optoelectronic phase shifter (OEPS). The transfer function of the phase shifter versus control voltage is linear with a maximum phase-tuning range of up to 1.95 π. The tuning responsivity of the OEPS is adjustable from 50 to 90 deg./V. The fluctuation and drift in phase of the controlled signal are approximately 0.050 and 0.003 deg./minimum.Timing jitter of the optical clock is less than 5 ps. Using a high-precision voltage regulator yields a tuning resolution of 0.2 deg. for 3-mV increments of control voltage. See "A Novel Frequency Discriminated Phase Controller for High-Speed Optical Clocks Generated From Laser Diodes," *IEEE Transactions On Instrumentation and Measurement,* April 2002, Vol. 51, No. 2, pp. 252-258.

Clock-Distribution Interconnect Features Rx, Tx, And Antenna

A WIRELESS INTERCONNECT system that transmits and receives RF signals across a chip using integrated antennas, receivers (Rxs), and transmitters (Txs) is examined by Brian A. Floyd of IBM T.J. Watson Research Center (Yorktown Heights, NY), Chih-Ming Hung of Texas Instruments, Inc. (Dallas TX), and Kenneth K. O of the University of Florida (Gainesville, FL). Using a 0.18-μm complementary-metal-oxide-semiconductor (CMOS) technology, each component is demonstrated at 15 GHz. Wireless interconnection for clock distribution is performed in two steps. First, a wireless Tx with integrated antenna generates and broadcasts across a 5.6-mm test chip. Then, a wireless Rx detects a 15-GHz global clock signal supplied to an on-chip transmitting antenna from 5.6-mm away from the Rx. See "Intra-Chip Wireless Interconnect for Clock Distribution Implemented With Integrated Antennas, Receivers, and Transmitters, *IEEE Journal Of Solid-State Circuits,* May 2002, Vol. 37, No. 5, pp. 543-552.

Recognize Target Type Using 1D Range Profiles

THE RADAR CROSS-SECTION of a target is dependent on operating frequency and angle, making the process of radar-target recognition a difficult chore. This is accomplished, however, through a technique that recognizes target type using 1D range profiles. According to Kyung-Tae Kim of the Dept. of Electrical Engineering and Computer Science, Yeungnam University (Kynongsan, Kyungbuk, Korea), Ph.D. candidate Dong-Kyu Seo, and Hyo-Tae Kim of the Pohang University of Science and Technology (Pohang, Korea), this technique uses the Multiple Signal Classification (MUSIC) algorithm to generate superresolved range profiles.

The central moments are computed and mapped into values between zero and unity, followed by a component analysis to delete feature-vector repetition. The retrieved features are classified according to the Bayes classifier. Results are obtained using five aircraft models measured in a compact range. These are compared with results obtained through conventional range profiles that are gathered by inverse Fast Fourier Transform. See "Efficient Radar Target Recognition Using the MUSIC Algorithm and Invariant Features," *IEEE Transactions On Antennas And Propagation,* March 2002, Vol. 50, No. 3, pp. 325-337.

Design Microstrip Bandpass Filters

MICROSTRIP BANDPASS FILTERS are found in microwave systems and commonly used as front-end filters in a receiver (Rx) or as an oscillator output filter in a transmitter (Tx). John R. Crute of Toracomm Ltd. (Bristol, England) describes a design procedure for miniaturized microstrip bandpass filters on high-ϵ_r substrates.

Electromagnetic (EM) simulations provide very close agreement with measured data. Crute confirms the procedure by designing a miniature microstrip hairpin-coupled bandpass filter at 1.8 GHz on $BaO–PbO–Nd_2O_3–TiO_2$ ceramic substrate. His simulation predicts a bandpass center frequency of 1840 MHz and a 3-dB bandwidth of 104 MHz compared to measured values of 1846 MHz and 99 MHz, respectively. Passband insertion loss is 2.5 dB versus a measured value of 3.1 dB. Simulated maximum passband return loss is 10.5 dB, whereas the measured value is 8.0 dB. See "CAD of High-ϵ_r Microstrip Bandpass Filters," *International Journal of RF and Microwave Computer-Aided Engineering,* May 2002, Vol. 12, No. 3, pp. 229-235.

Tracking Advances In VCO Technology

Improvements in monolithic RF IC VCO technology have paved the way for increased use of integrated sources in high-volume commercial communications applications.

Voltage-controlled oscillators (VCOs) are commonly found in wireless systems and other communications systems that must tune across a band of frequencies. VCOs are available from a wide range of manufacturers in a variety of package styles and performance levels. But modern surface-mount and RF-integrated-circuit (RF IC) VCOs owe their heritage to engineering developments that began approximately a hundred years ago. Improvements in VCO technology have continued throughout that time, yielding ever-smaller sources with enhanced phase noise and tuning linearity.

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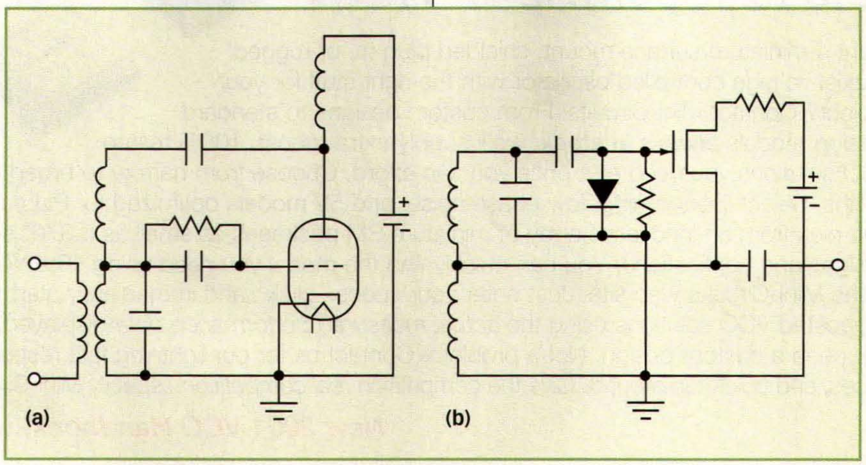
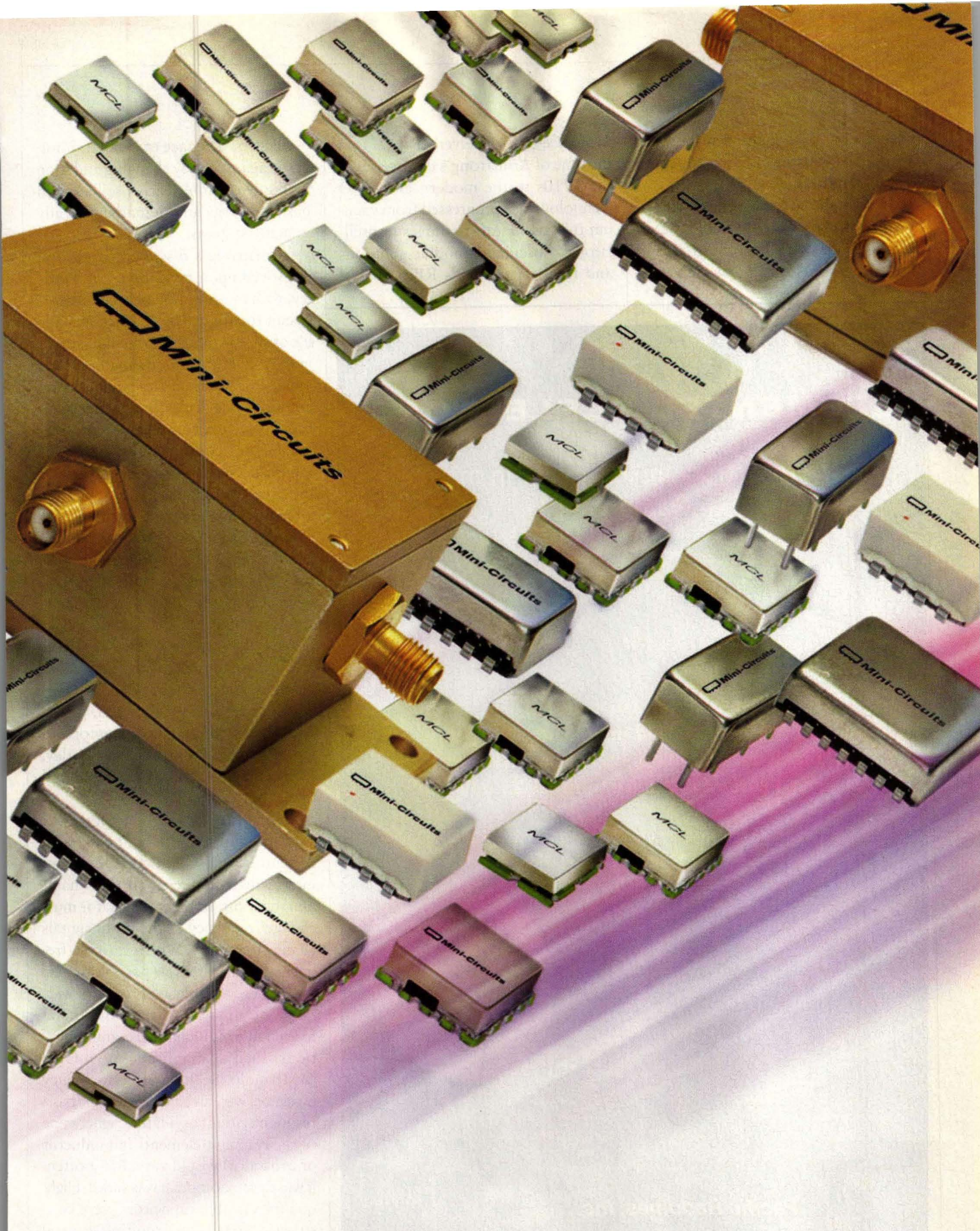
Oscillators have been essential components from the time Edwin Armstrong discovered the heterodyne principle.[1] In this application, an oscillator feeds sinusoidal signals to a nonlinear mixing element to effect frequency translation by multiplying the oscillator's signals with other input signals. Of course, Armstrong realized that what he needed to control the frequency translation was an electrical circuit which produced a stable sinusoidal time-varying voltage (or current) with a corresponding frequency. He discovered around that same time that an Audion (an early vacuum tube) could be configured to produce an oscillation, and he effectively devised the first



1. These examples of the Hartley oscillator show the triode implementation (a) and the JFET implementation (b).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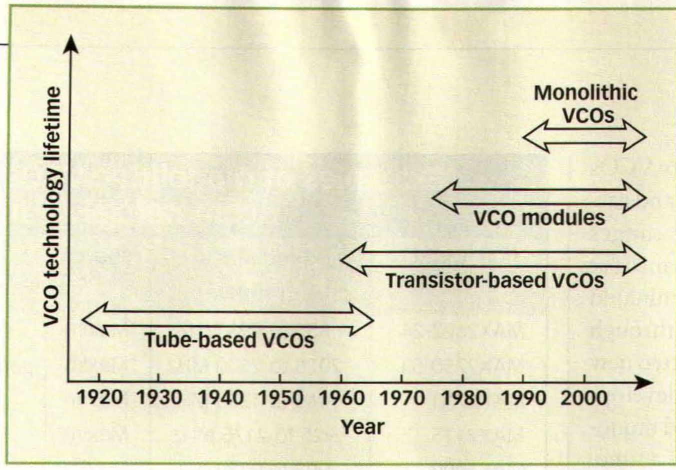
electronic oscillator[2] [rather than the crude spark-gap oscillators used in early wireless transmitters (Txs)].

In retrospect, Armstrong started a revolution in oscillator technology that quickly made spark Txs obsolete, leading to the development of high-performance radio receivers (Rxs). From the time of Armstrong's discoveries in the 1910s to the modern era, VCO technology has progressed from vacuum-tube oscillators to transistor oscillators to oscillator-module solutions and finally to today's RF IC-based oscillators. The face of VCO technology is again rapidly changing and soon will only resemble early oscillators in basic topology and/or mathematically in many systems.

Armstrong's discovery was soon improved upon by Robert V.L. Hartley, with the invention of his oscillator-circuit topology **(Fig. 1)**. Hartley made use of improvements in vacuum-tube technology and devised a oscillator circuit where the vacuum tube acted as an amplifying device with inductive feedback applied to create a regenerative oscillation. The frequency of oscillation was established by the coil inductance and the circuit capacitance. This circuit was a breakthrough in the generation of a sinusoidal signal—it provided a much greater range of possible frequencies simply by varying the value of the coil or capacitor. The Hartley oscillator circuit was popular in Txs and was quickly adapted for use in World War I. Txs and Rxs made use of the new tube-based oscillators circuit. Oscillator circuit innovations proliferated, giving rise to the predominant circuit topologies still in use today, such as Hartley, Colpitts, Clapp, Armstrong, Pierce, and other topologies.

In Armstrong's superheterodyne Rx principle, input signals are mixed with oscillator signals to produce a constant intermediate frequency (IF). To maintain the constant IF, the oscillator must change frequency as the input signals change frequency. With a variable-frequency oscillator it was possible to tune the frequency-translation circuit to a wide range of input RF signals and therefore enable multichannel communications, such as amplitude-modulated (AM) radio. Such variable-frequency oscillators were an adaptation of the basic resonant-circuit oscillators, where one of the resonant elements (an inductor or capacitor) would vary. Most often, it was the capacitor that was varied. High-quality variable capacitors were constructed from ganged multiplate metal air-gap capacitors. As radio technologies advanced, a tremendous amount of innovation took place in the implementation of oscillator circuits. Engi-

neers devised countless types of coils, variable capacitors, feedback techniques, and vacuum tubes to implement oscillator and frequency-conversion circuits. Many elaborate and elegant schemes were devised to provide precise, high-quality tuning of the oscillator frequency through a mechanical dial on the front on the radio. As with many early implementations of electronics, these circuits were bulky and expensive and required high supply voltages.



2. This chart shows changes in VCO technology lifetimes as a function of time.

The vacuum-tube oscillator was widely employed for many years in commercial and military-radio Rx applications, such as AM and frequency-modulated (FM) radios, television, and military voice communications. However, the discovery of semiconductor amplifying devices, such as the transistor and the varactor diode, led to the next dramatic change in VCO technology. The first bipolar transistor was discovered in the late 1940s at Bell Laboratories (Holmdel, NJ), and transistors became available in the 1950s as replacements for vacuum tubes. The new transistors were smaller and consumed less power than tubes, with lower operating voltage requirements and ultimately lower cost. The transistor became a replacement for the vacuum tube as the active element in oscillators and significantly changed the practical implementation established oscillator topologies.

Arguably, the introduction of the varactor diode (with a voltage-variable capacitance arising from a reverse-biased PN junction) had a greater impact on the direction of VCOs than the transistor. In the early 1960s, a great deal of research was performed on varactor technology, and varactors rapidly displaced mechanically adjustable components as the variable-capacitance element in VCOs. Varactors proved invaluable in the development of phase-locked-loop (PLL) circuits for precise electronic control of frequency sources. The rapid growth of television during that time contributed greatly to the migration to varactor- and transistor-based VCOs. Cost-effec-

tive, low-power, and high-quality VCOs with inherent electronic tuning and easily reconfigurable frequency ranges were now possible. Discrete-transistor and varactor-based VCOs dominated electronic designs of the 1960s through 1980s. In the 1980s, however, two new technologies impacted VCO developments: modular approaches and monolithic VCO ICs. **Figure 2** shows a timeline illustrating the development of VCO technologies over the past 80 years.

The shrinking sizes of varactors, capacitors, and inductors made VCOs in module form possible. A VCO module is essentially a miniature version of a discrete-component oscillator constructed on a substrate that is mounted into a metal housing. The module is self-contained and requires only connections to ground, the supply voltage, the tuning voltage, and the output load. These modules first appeared in the 1960s, primarily for military applications. They were fairly large (several square inches) and relatively expensive. Discrete transistor and varactor implementations of VCOs were still used in commercial products. It was not until the emergence of mobile telephony that a commercial market emerged for VCO modules.

Although discrete VCOs could be custom designed to any frequency and tuning range, they typically required labor-intensive production adjustment of the frequency-setting elements to compensate for component variations. In addition, discrete VCOs needed good shielding to minimize emissions and reduce pulling effects. With the growing sales of mobile telephones in the late 1980s and early 1990s, demand increased for "canned" oscillator modules. Some Japanese companies, being increasingly proficient in miniaturization, developed small, cost-effective VCO modules for mobile telephones. As new, wireless applications arose, VCO-module manufacturers developed products with frequency plans unique to each application. As surface-mount components became progressively smaller (1206, 0805, 0603, 0402, and 0201), new VCO modules were developed these components, creating smaller, lower-cost modules. **Figure 3** illustrates the size reduction over time of the "typical" state-of-the-art commercial VCO module. Today, these improvements have culminated in compact (4 × 5 × 2-mm) modules that sell for close to $1.00 (US) in high volumes. This 15-year cycle of shrinking VCO-module volume was a truly amazing reduction in size and satisfied the tough space constraints imposed by the new mobile wireless devices, such as cellular phones. Yet an even smaller and more cost-effective VCO technology would emerge by the end of the 1990s—monolithic VCO-IC technology.

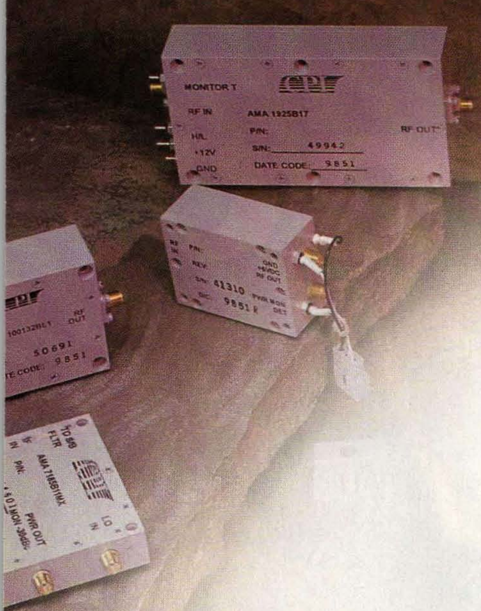
Monolithic VCO-IC technology is defined as a VCO implementation in which all the circuit elements of an inductive-capacitive (LC) VCO—transistors, capacitors, resistors, inductors, and varactor diodes—are integrated on one chip. As in a VCO module, the devices are configured to form a complete VCO, requiring only connection to the power supply, ground, output, tuning input, and any digital control lines. (Note that voltage-controlled ring-oscillator circuits have been excluded from this definition of VCOs, given that their phase noise is much poorer and eliminates its use in most radio systems.) The first instance of a monolithic VCO IC coincided with the development of gallium-arsenide (GaAs) IC technology and monolithic-microwave ICs (MMICs). The monolithic VCO emerged in the literature[A,B] in the early 1980s during a period of intense research into commercial and military applications for MMICs (funded largely by the US DARPA MIMIC program). Early MMIC VCOs were fabricated with GaAs IC processes, using 2-in. (5.08-cm)-diameter wafers, although the MMIC VCOs were not particularly area efficient and, therefore, were not cost effective. Generally, these VCOs operated at multigigahertz frequencies that were consistent with the target applications, satellite Rxs and radar systems.

Most of the early monolithic GaAs VCOs were developed as part of the DARPA MIMIC research, with little impact on commercial markets. Silicon (Si)-IC technology was still relegated to low frequencies during the 1980s, and lacked the bandwidth needed for gigahertz-frequency monolithic VCOs. But by the 1990s, Si-IC technology had been developed with sufficiently high transition frequencies (f^T) and suitable monolithic components [high-quality-factor (Q) inductors and high-frequency capacitors and varactor diodes] to enable development of higher-frequency Si monolithic VCOs. Wireless markets had emerged with sufficient

Examples of monolithic VCOs integration in commercial RF ICs

Unit	Frequency range	Source	Application
MAX2622-24	855 to 998 MHz	Maxim	general purpose 900-MHz ISM
MAX2750-53	2025 to 2500 MHz	Maxim	general purpose 2.4-GHz ISM band
MAX2754	1145 to 1250 MHz	Maxim	2.4-GHz cordless phones
MAX2115	925 to 2175 MHz	Maxim	DBS
MAX2900	902 to 928 MHz	Maxim	900-MHz ISM band (wireless meter reading)
MAX2820	2400 to 2500 MHz	Maxim	802.11b WLAN
RF105	902 to 928 MHz	Conexant	900-MHz cordless phones
SA2400	2400 to 2500 MHz	Philips	802.11b WLAN
BlueCore-01	2400 to 2500 MHz	CSR	Bluetooth
TRF	2400 to 2500 MHz	TI	Bluetooth
GRF2i/LP	1575 MHz	SiRF	GPS
AR5111	5.2 to 5.8 GHz	Atheros	802.11a WLAN

Amplifiers



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Broadband, Small Signal

Model Number	Frequency Range (Ghz)	Gain (dB Min)	Gain Flatness (±dB Max)	Noise Figure (dB Max)	VSWR Input Port Max	VSWR Output Port Max	Output Power @ 1dB CP (dBm Min)	DC Input Current Vdc:+12 (mA Typ)
CMA2080A1	2.0-8.0	30	1.5	6	2:1	2:1	+15	200
CMA20120A	2.0-12.0	33	2.0	6	2:1	2:1	+15	350
CMA20180A	2.0-18.0	34	2.0	6	2:1	2:1	+18	450
CMA60180A1	6.0-18.0	36	1.5	6	2:1	2:1	+15	350
CMA180265A	18.0-26.5	30	1.5	6	2:1	2:1	+16	400
CMA265400A	26.5-40.0	30	1.5	6	2:1	2:1	+16	400

Broadband, Low Noise

CMA60180A2	6.0-18.0	30	1.5	3	2:1	2:1	+10	200
CMA180265A1	18.0-26.5	30	1	3	2:1	2:1	+10	200
CMA265400A1	26.5-40.0	28	1.5	3.5	2:1	2:1	+10	200

Medium Power

CMA5964B10	5.9-6.4	40	1.0	8	1.5:1	1.5:1	+33	1500
CMA5971B1	5.9-7.1	20	1.0	10	1.8:1	1.8:1	+33	1500
CMA7185B2	7.1-8.5	20	1.0	10	1.8:1	1.8:1	+33	1500
CMA85125B1	8.5-12.5	30	1.5	8	2:1	2:1	+35	3000
CMA107117B2	10.7-11.7	20	1.0	10	1.8:1	1.8:1	+33	2000
CMA127132B	12.7-13.2	40	1.0	5	1.8:1	1.8:1	+34	4000
CMA137145B	13.7-14.5	45	1.0	6	1.5:1	1.8:1	+33	1500
CMA142153B6	14.2-15.3	15	1.0	8	1.5:1	1.8:1	+30	1000
CMA177197B15	17.7-19.7	35	1.0	8	1.5:1	2:1	+30	1100
CMA181186B17	18.1-18.6	34	0.5	10	1.5:1	1.5:1	+33	3000
CMA200230B1	20.0-23.0	10	1.0	12	1.5:1	2:1	+30	1000
CMA295297B1	29.5-29.7	20	0.3	10	1.5:1	1.8:1	+30	1000

High Power

CMA1616B	1.6-1.68	45	0.25	10	2:1	2:1	+43	8500
CMA4450B27	4.4-5.0	40	1.0	8	1.5:1	1.5:1	+43	11000
CMA5964B40	5.9-6.4	40	1.0	8	1.5:1	1.5:1	+43	12000
CMA127132B7	12.7-13.2	40	1.0	8	1.5:1	1.5:1	+43	20000
CMA137145B19	13.7-14.5	53	1.0	6	1.5:1	1.5:1	+43	22000

TWT/KPA Drivers, Linearized Gain Control

Model Number	Frequency Range (Ghz)	Gain (dB Min)	Gain Flatness (±dB Max)	Noise Figure (dB Max) @ 0 Gain Control	VSWR In/Out Max	Gain Control (dB Max)	Output Power @ 1dB CP (dBm Min)	DC Input Current Vdc:+12 (mA Typ)
CMA5866A13	5.8-6.6	30	1.0	7	1.4:1/1.3:1	25	+13	260
CMA7984A1	7.9-8.4	30	1.0	7	1.4:1/1.3:1	25	+13	260
CMA127145A6	12.7-14.5	35	1.5	7	1.4:1/1.3:1	25	+18	500
CMA173184A8	17.3-18.4	38	1.0	7	1.4:1/1.3:1	25	+20	500
CMA270310A4W/G	27.0-31.0	20	1.0	10	1.5:1/2.0:1	25	+20	500

Note: Gain control voltage range is 0 to +10 Vdc (Maximum gain @ +10 Vdc)

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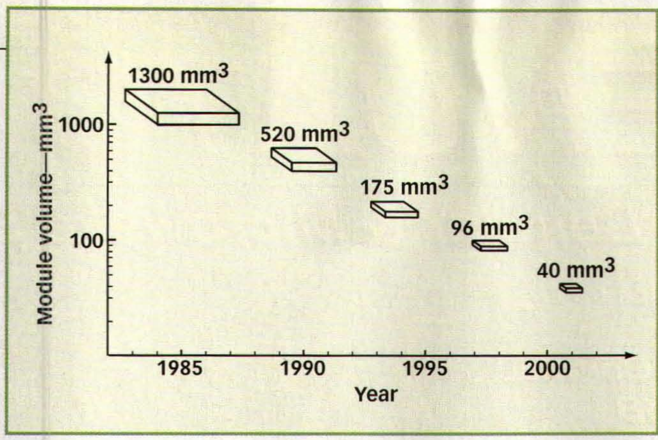
size and growth potential to spur the demand for low-cost VCOs in the 800-to-2500-MHz bands.

Prior to this, most commercial radio systems operated at frequencies low enough to make construction of a monolithic VCO IC impractical. On-chip inductor values were simply too large. The first apparent instance of a Si monolithic VCO IC in the literature is from the University of California at Berkeley in 1992.[3] The VCO employed a unique, unorthodox topology where the frequency was varied by electrically "interpolating" between two separate resonant circuits, even though it was still technically an implementation of monolithic Si VCO-IC technology. Arguably, this work and further research by Professor Robert Meyer and his graduate students at the University of California at Berkeley appears to have ushered in a period of increased research on monolithic VCOs.



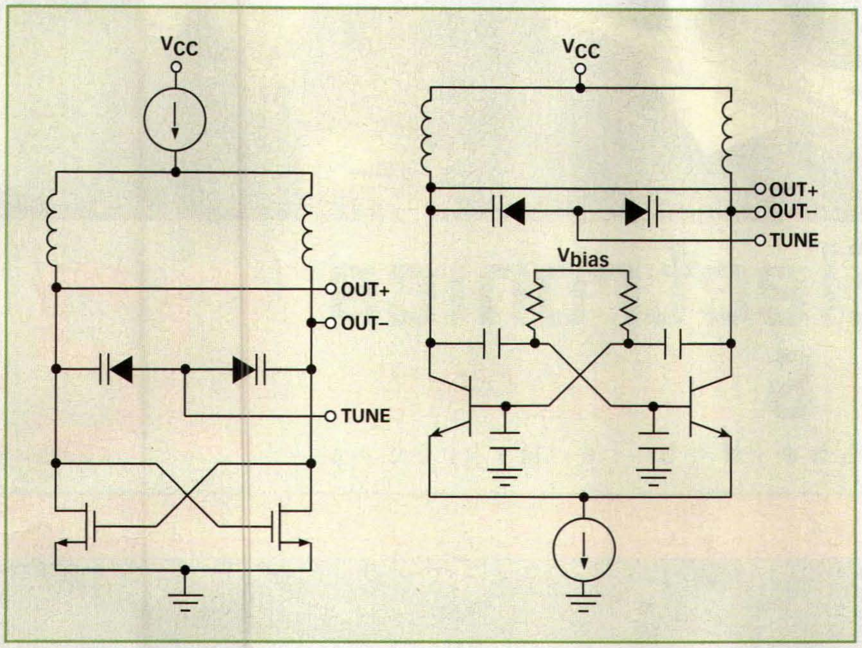
3. This plot shows VCO module size scaling as a function of time.

By 1995, work on Si monolithic VCO ICs was being reported in the technical literature by researchers at leading universities.[4,5] In these reports, researchers disclosed some of the first examples of modern, monolithic LC resonator VCO ICs. From 1996 to 1997, a tremendous number of papers appeared describing work on different implementations of monolithic VCOs.[6-12] This period effectively marked the emergence of the commercially viable monolithic VCO ICs. The monolithic VCO ICs were being developed in high-frequency bipolar-transistor-IC technology and Si complementary-metal-oxide-semiconductor (CMOS)-IC technology. Academic researchers typically used CMOS technologies to take advantage of the widespread availability of the IC technology, while industrial researchers used RF IC-specific bipolar-CMOS (BiCMOS) process technology. **Figure 4** shows a typical monolithic VCO circuit implemented in CMOS and BiCMOS process technology.

Generally, the overall performance of these early VCO-IC implementations was inferior to discrete implementations and VCO modules. Specifically, the phase noise and tuning characteristics were poorer than what could be routinely achieved in discrete designs or VCO modules. This shortfall was principally due to the low-Q inductors and crude varactor diodes commonly available in that generation of IC technologies.

However, monolithic VCOs proved to be extremely small, cost-effective, and available in the same process where RF transceiver functions were being implemented. This meant that the VCO could be integrated with other RF and IF functions, such as the mixer, low-noise amplifier (LNA), and PLL. This capability to cost-effectively integrate the VCO with other Rx and Tx functions helped make the monolithic VCO IC a commercial reality. A good early example of this was a commercial 900-MHz spread-spectrum cordless-telephone chip set.[13]

In the late 1990s, research on VCO-IC technology intensified considerably.[14-20] This was in large part due to the explosion in the wireless markets and also in the proliferation of high-frequency bipolar, CMOS, and BiCMOS process technologies. Significant research and development took place at industrial and academic levels. Researchers focused on improving the phase-noise performance, extending the frequency of operation and adjustment of the VCO's tuning range on-chip. These improvements achieved electrical specifications which permitted the VCOs to be used in RF ICs for cordless phones,



4. These typical monolithic VCO core circuits are shown in MOS and bipolar forms.

Bluetooth, wireless-local-area-network (WLAN), Global Positioning System (GPS), and direct-broadcast-satellite (DBS) applications. **The table** shows a summary of some commercial RF ICs which contain monolithic VCOs.

These VCO ICs and the integrated solutions that contain them are smaller and more cost-effective than VCO modules and easier and faster to apply than discrete solutions. The monolithic VCOs provide significantly improved value over previous technologies. The performance of this generation of VCO technology is sufficient for systems such as cordless phones, wireless data radios, and DBS Rxs, and, therefore, is being widely adopted for use in these systems. However, the phase-noise performance is presently insufficient (the noise is approximately 5 to 10 dB too high) to meet the requirements of higher-date-rate mobile telephone systems [such as Global System for Mobile Communications (GSM), IS-136, code-division multiple access (CDMA), etc]. Low inductor Q and excess bias noise contribute to limits for the VCO phase noise. Although some researchers had demonstrated promising results with the use of bond-wire inductors, low phase-noise performance has remained elusive and out of reach of monolithic VCO-IC technology. However, this is appears to be only temporary. In the last three years, many significant advances in VCO design have been reported and point out some clear trends for the future.

Major Trends

Several trends are impacting the development of monolithic VCOs with improved phase noise. For example, basic RF IC process technologies are improving. The Qs possible with semiconductor processes is increasing, and the performance of active and passive devices is improving. Even with Si processes, transistors can now be fabricated with f^T performance exceeding 50 GHz, and higher-Q varactor diodes are available with wide capacitance ratio tuning ranges (low series resistance). These processes feature lower-loss substrates with thicker metalizations and higher-Q inductors. The processes are capable of devices with reduced parasitic elements, leading to VCOs with lower phase noise, higher operating frequencies, and lower-current operation.

Design techniques are also becoming more advanced. VCO researchers are exploiting the power of IC technologies by devising more sophisticat-

ed circuits to improve performance. These researchers are introducing techniques previously impractical with discrete VCO or module VCO implementations, such as differential oscillator topologies, amplitude control, second-harmonic traps, IC transformers for improved coupling, topologies with multiple oscillators, and architectures capable of higher-frequency operation.

Design engineers are also gaining a better understanding of VCO theory. They are building upon mathematical models from the past, such as Van der Pol's and Leeson's equations, and devising new analytic expressions for oscillator behavior (such as tuning characteristics and phase-noise performance). For example, designers are on the cusp of amending Leeson's noise equations with Abidi's relationships. In addition, computer-aided-engineering (CAE) tools are growing in power and sophistication as the processing capabilities of workstation and personal computers (PCs) increase, allowing engineers to experiment with VCO behavioral models to discover performance enhancements.

Monolithic VCO technology continues to appear in an increasing number of new products, with high-quality VCOs integrated with the transceiver circuitry. For example, the latest transceivers for the WLAN and Bluetooth markets integrate the VCO within the RF transceiver IC, resulting in a dramatic reduction in size compared to discrete components. In higher-performance WLAN radios (2.4-GHz 802.11b and 5-GHz 802.11a versions), system requirements call for higher-performance VCOs with the very-low phase noise needed to achieve the required packet data rates and blocking performance levels.

Improvements in RF IC VCO technology make these integrated sources ever more attractive for an increasing number of commercial RF applications, including satellite Rxs, cable-television (CATV) set-top boxes, wireless data applications, cordless telephones, and mobile telephones. Clearly, monolithic VCOs are winning an ever-increasing share of high-volume applications compared to discrete and module VCO solutions. The time is coming very soon when monolithic VCOs will be the dominant oscillator approach in all high-volume commercial wireless systems. VCOs have traversed a remarkable path from bulky tube based circuits to less than $1 mm^2$ of Si. MRF

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FOOTNOTES

A. The heterodyne principle is defined as the multiplication of two signals in the time domain to produce a frequency shift in the frequency domain. The principle is the fundamental basis for frequency translation of signals in wireless systems.

B. Both Edwin Armstrong and Lee DeForest were working on regenerative Rx circuits at the time. These regenerative circuits created the first oscillators.

Setting Bias Points For Linear RF Amplifiers

The choice of biasing arrangement can determine the ultimate performance of a wireless PA in terms of output power, efficiency, linearity, and other parameters.

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Choosing the bias points of an RF power amplifier (PA) can also determine the level of performance ultimately possible with that PA. By comparing PA bias approaches, designers can evaluate the trade-offs of the approaches when used for different applications. Last month, we studied the effect of using a constant reference voltage (a regulated supply sets the bias point), and using bias-current feedback (where a feedback loop maintains a constant bias current). This month, we will examine dynamic bias configurations where the RF output power is used to set the bias point. In the first example, the bias current is set through the output power, P_{out} (where the RF output power adjusts the bias current). This approach will then be expanded to having the bias current and voltage set through P_{out} (where the RF output power sets the PA bias current and voltage).

The constant V_{reg} bias approach can be modified to demonstrate how the output power may be used to adjust the bias point **(Fig. 10)**. The bias-current-feedback technique, discussed last month, could be modified to produce the same performance, but the implementation is easier to visualize with the constant V_{reg} technique.

The power-sense voltage is derived by the same method presented last month, with the only change being that the opamp2 drives an additional voltage-to-current converter (implemented using the 30-kΩ resistor), which drives the n-channel current mirror. Adjusting the size of the 30-kΩ resistor provides scaling of the transfer function.

The modification to the constant V_{reg} bias approach is in the additional 120-kΩ resistor and shunting-current mirror added to the feedback path of opamp1. To visualize the operation of opamp1, assume the output power is high and the VPS (output of opamp2) drives the n-channel current mirror hard enough to short the 120-kΩ resistor. Opamp1 now operates like the previously demonstrated constant V_{reg} approach (compare to Fig. 3 last month) and provides approximately +2.7 VDC for the V_{reg} input on the RF5117. Now assume that the output RF power is zero. The n-channel current mirror driven by opamp2 is off, and opamp1 generates a much lower V_{reg} voltage due to the larger resistor (120 kΩ + 32 kΩ) from its inverting input to ground. Voltage V_{reg} without RF applied is approximately +1.42 VDC. This V_{reg} sets a quiescent current in the RF5117 PA of 17.3 mA.

Figures 11 and 12 show the RF and linearity performance using this biasing approach, with extremely nonlin-

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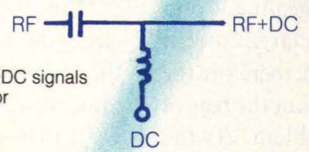

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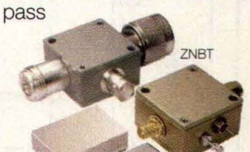

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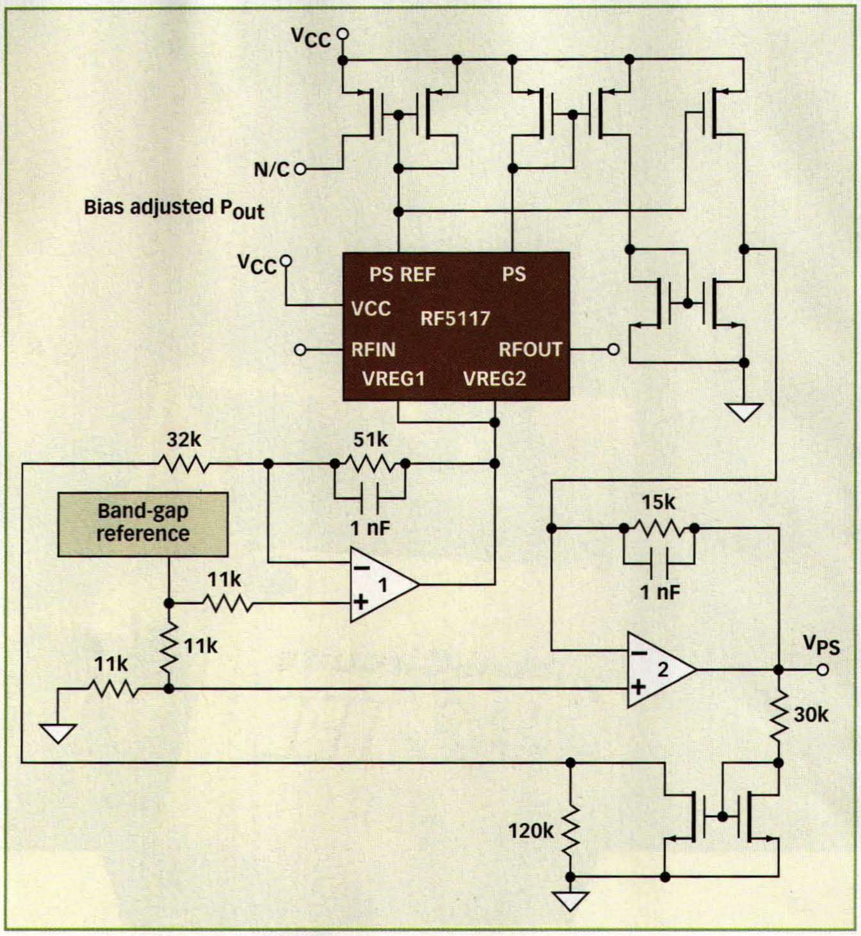
ear gain apparent. This is a characteristic of the heterojunction-bipolar-transistors (HBTs) where the beta is dependent on the current density. This "extreme" gain expansion may be unexpected, but it is not detrimental to the amplifier performance. The input RF power is changed due to a desire to modify the output power. The gain expansion in the PA reduces the dynamic range needed in a variable-gain amplifier. Also, a measure of the output power is being provided by opamp2 (VPS in Fig. 10). In applications where a well-defined PA output power is required, a feedback loop controlling a variable-gain amplifier may be closed around the PA. The loop will set the PA output power while compensating for gain variations due to device variations and temperature.

Figure 12 shows the effect of the gain expansion on linearity. The adjacent-channel-power-ratio (ACPR) goal of at least –33 dBc is being satisfied. The RF5117 is being maintained under Class AB bias conditions even at reduced output powers. The lowpass filtering in the bias-generating loops prevents the AM-to-AM distortion expected from this much gain expansion.

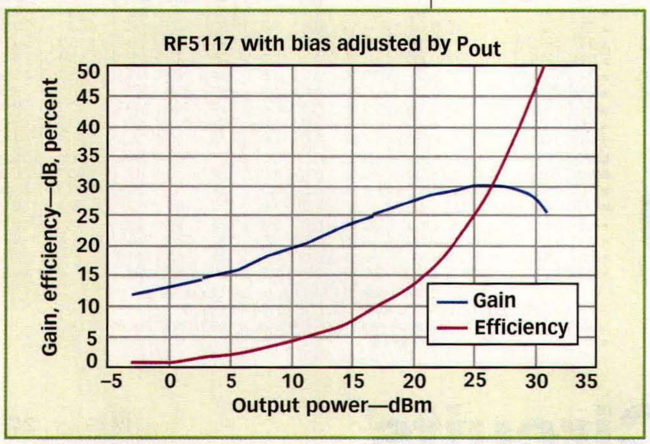
This approach provides an improvement in efficiency at reduced output power. If the bias point can be cut back at low output powers, this bias approach can provide a reduction in battery drain. To put numbers on this improvement, the battery drain with the constant V_{reg} bias at 0-dBm output power is 137 mA and the battery drain using P_{out} to adjust the bias is 26.9 mA at 0-dBm output power.



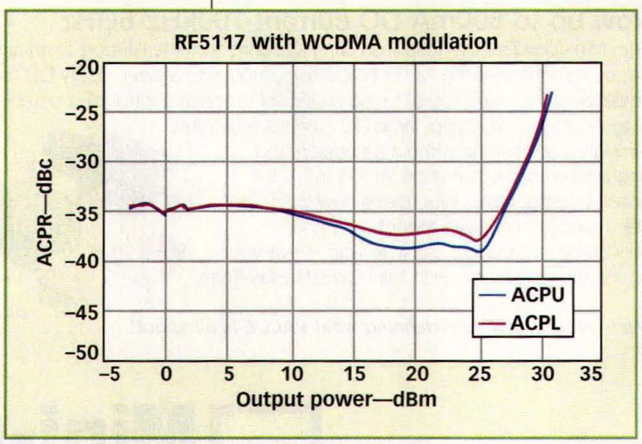
10. RF5117 bias adjusted through P_{out} is shown here.

Figure 13 shows how the control loops adjust the bias voltage. At low output power, there is not a large current swing in the output device, therefore the bias can be reduced without the fear of causing a linearity problem. As the power increases, V_{reg} increases to +2.7 VDC as the current mirror saturates across the 120-kΩ resistor in Fig. 10.

The next step in improving the efficiency during power backoff is to reduce the V_{cc} voltage applied to the PA. The improvement can only be realized if the DC-to-DC converter can transform the



11. RF5117 performance using a V_{reg} bias adjusted through P_{out} can be seen here.



12. This shows RF5117 linearity performance using a V_{reg} bias adjusted through P_{out}.

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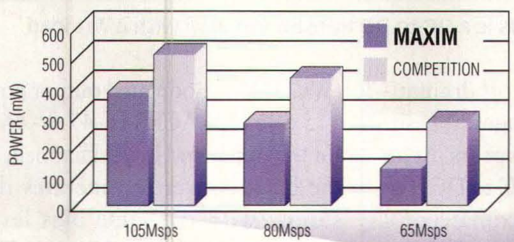

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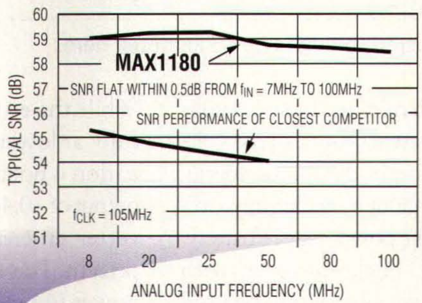

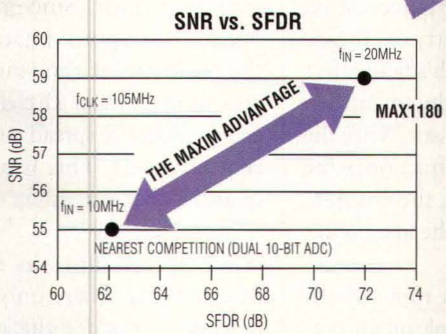

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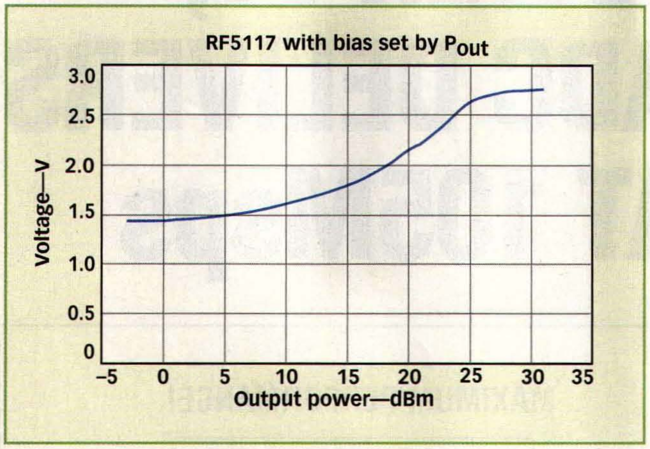
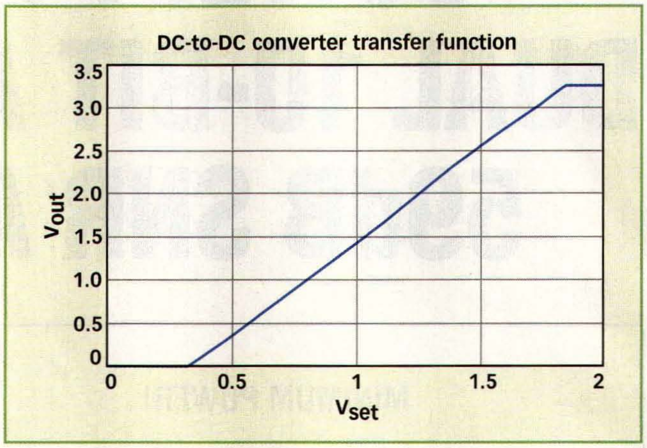


13. V_{reg} bias adjusted through P_{out} is illustrated here.



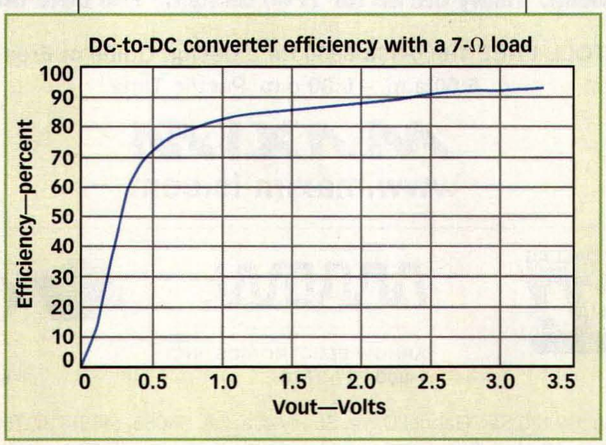
14. This is a DC-to-DC transfer function with a 7-Ω load.

voltage level with minimal losses. A voltage-dropping resistor or series-pass voltage regulator simply dissipates energy outside the PA without providing improvement as seen by the power supply/battery.

A DC-to-DC buck converter with an analog voltage control will be used for this purpose. This function can be thought of as a box that takes in +3.5 VDC and turns out lower voltages as determined by a V_{SET} control signal. **Figure 14** shows the transfer function of the converter with a 7-Ω resistive load. The load was chosen to approximate the power-supply drain of the RF5117 when it is supplying approximately +27-dBm RF power.

The output of the DC-to-DC converter is zero until V_{SET} reaches approximately +0.35 VDC. It then linearly increases until V_{SET} reaches approximately +1.8 VDC, where the output saturates at about +3.27 VDC. The +0.23-VDC difference between the input supply and the maximum output voltage is accounted for in the voltage drop across the internal-switch transistor and the inductor.

When the characterization data was recorded for the DC-to-DC converter, the input and output current and voltage were monitored, allowing a calculation of the converter efficiency. (The resistance was not used since it changes slightly as power increases.) **Figure 15** shows the calculated efficiency of the converter with the 7-Ω load resistor.



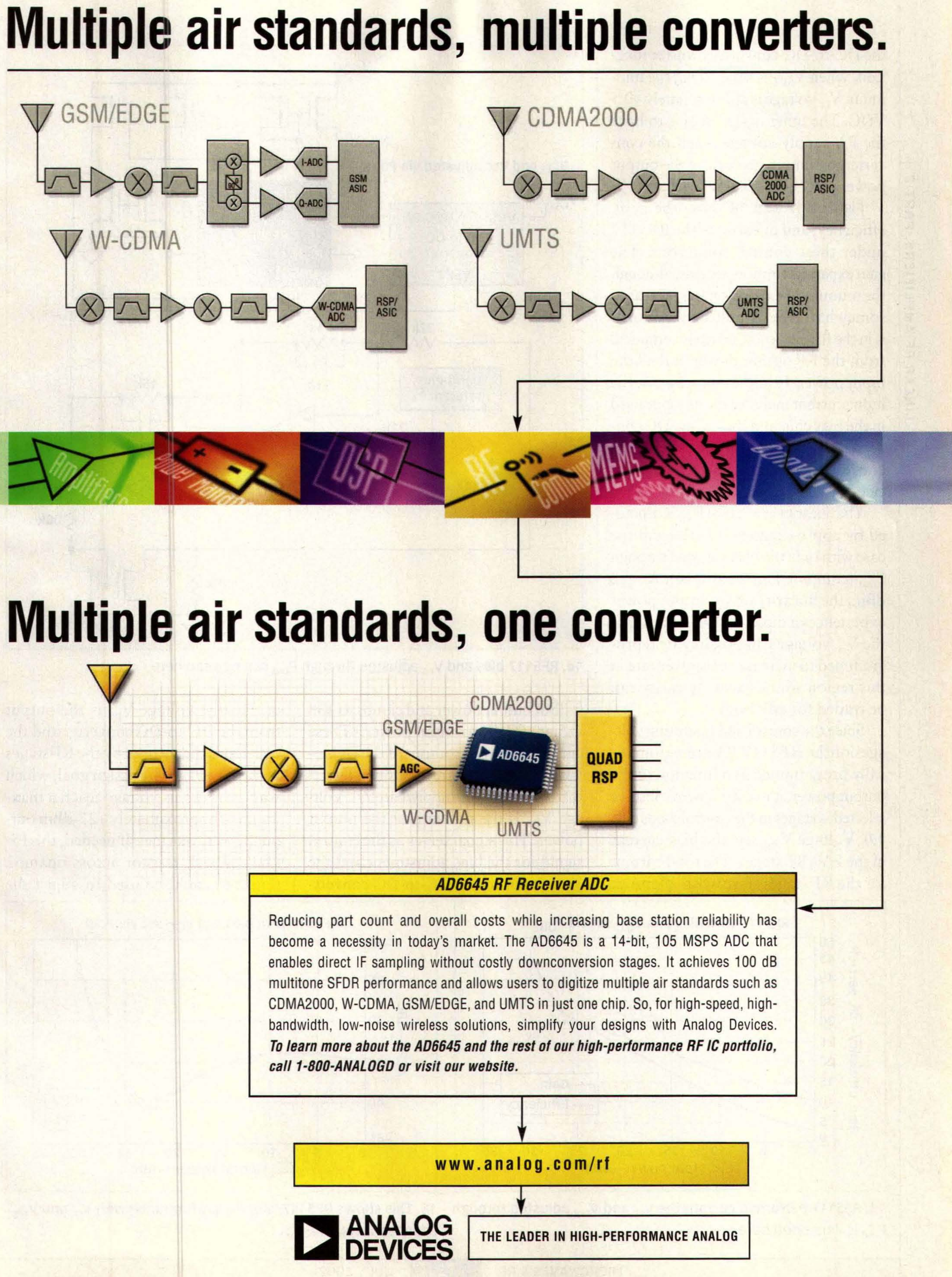
15. DC-to-DC efficiency is shown with a 7-Ω load..

While the efficiency trails off dramatically at lower output voltages, this situation is not as serious it appears. At an output of +0.46 VDC, the DC-to-DC converter efficiency is approximately 70 percent. This is computed from the input power (44 mW) and the output power (31 mW). The lost power is accounted for in the controller circuitry overhead, resistive losses in the switch's field-effect transistors (FETs) and inductor, and in the capacitive switching losses. While the ratio indicates that a large fraction of the input energy is being lost, the absolute quantity of 13 mW puts the lost power in a better perspective.

A potential drawback to a switching power supply is the ripple on the output voltage. This can potentially mix with the signal passing through a PA and create undesirable spurious signals. When the RF5117 is biased at V_{reg} = +2.7 VDC, V_{cc} = +2.0 VDC with a wideband-code-division-multiple-access (WCDMA)-modulated output power of +15 dBm, the ACPR is –47.8/–46.3 dBc for the upper and lower channels. When the buck converter generates the V_{cc} supply with the same bias levels as before, the ACPR is –47.7/–46.1dBc. The differences are within the ACPR measurement error. Since the converter operates at approximately 500 kHz, the majority of the spurious energy appears in the main RF channel, but some energy should be spread out into the adjacent channels. This is an attempt to quantify that spreading energy.

Reducing the PA V_{cc} voltage at low power levels improves the efficiency over the case where only the bias was reduced. The added efficiency is derived from the step-down action of the DC-to-DC converter **(Fig. 16)**. The power-supply (battery) current is reduced by the step-down voltage ratio. (The non-ideal converter efficiency accounts for the step-down losses).

The addition of the DC-to-DC converter to power the RF5117 is the only change that was made to the previous case where the bias was adjusted through P_{out}. The output-power measurement V_{PS} becomes the controlling function for the V_{SET} input to the DC-to-DC converter. Two V_{PS} data points are important in setting up the control loop. At low output powers, the V_{PS} is approximately +0.6 VDC, and at +27 dBm (0.5 W) the V_{PS} is approximately +2.0 VDC. From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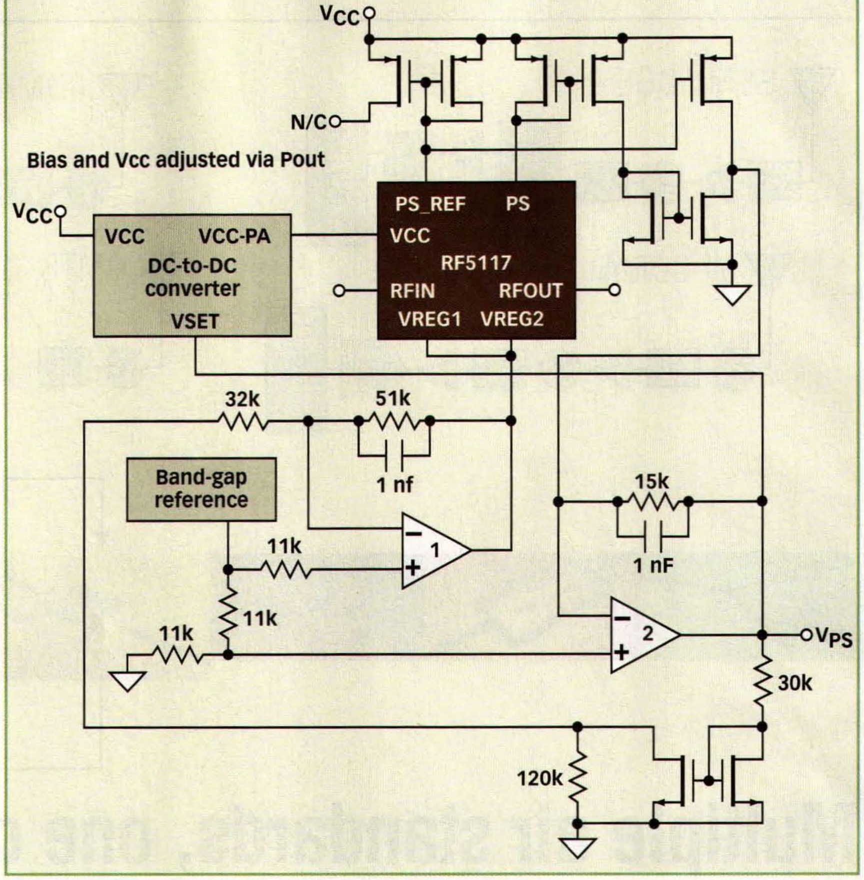


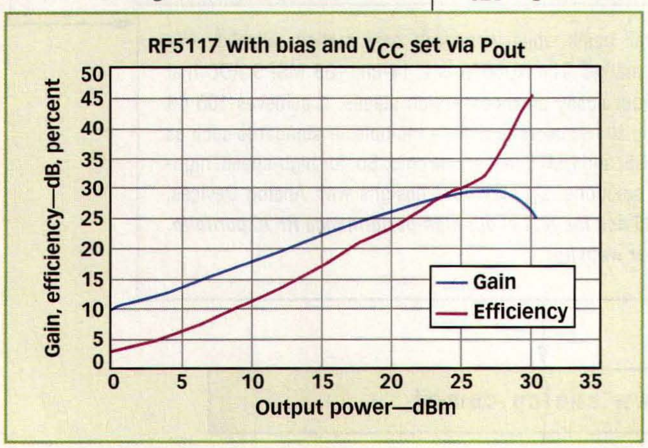
16. RF5117 bias and V_{cc} adjusted through P_{out} can be seen here.

the DC-to-DC-converter transfer function, when V_{SET} is +0.6 VDC, the minimum V_{cc} voltage is approximately +0.5 VDC. The other design goal is to have the PA supply voltage reach the converter maximum just before the output power reaches the 0.5-W level.

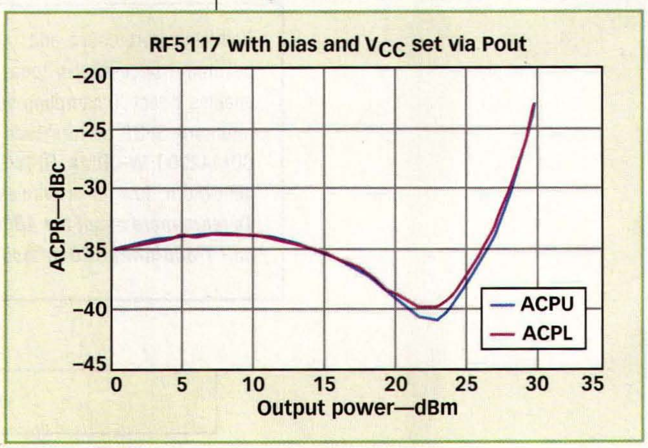
Figures 17 and 18 show the gain, efficiency, and linearity of the RF5117 under these control conditions. The gain expansion is more regular, although the amount of expansion has increased somewhat. The largest improvement is in the RF efficiency, which is computed from the RF output power and all the input power. In addition to the PA, the input current includes the amount used in the bias chip and the DC-to-DC converter. At 0-dBm output power, the entire system maintains a total efficiency of 3 percent.

The linearity at +7 dBm P_{out} is degraded by approximately 0.5 dB from the case with only the bias adjusted through P_{out} (compare Figs 12 and 16). At +22 dBm, the linearity of the output power experiences a dip, which indicates that the V_{cc} voltage can possibly be reprogrammed to increase at a slower rate in this region so the excess linearity can be traded for efficiency.

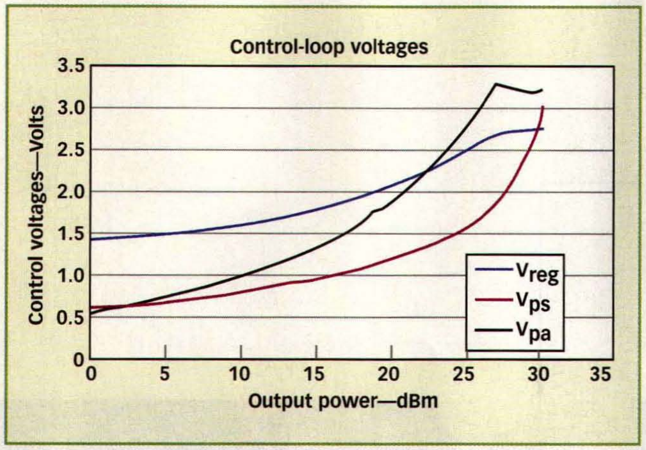
Since the control and operating voltages of the RF5117 PA are automatically programmed as a function of the output power, it is instructive to observe selected voltages in the control loops **(Fig. 19)**. Voltage V_{reg} sets the bias current in the PA's RF stages. The total current for the RF stages starts with 23 mA at 0-dBm output power and climbs to 438 mA at +27 dBm output power. (These currents are at the output of the DC-to-DC converter and are not drawn directly from the power supply/battery). Voltage V_{PS} is the measure of the output power. This signal serves as the control signal for the bias adjustment and the V_{SET} input to the DC-to-DC converter. Finally, voltage V_{pa} is the output from the DC-to-DC converter and the collector voltage used by the RF stages in the RF5117. The design goal, which was to have this voltage reach a maximum of approximately +27-dBm output power, was met. If needed, the 15-kΩ feedback resistor across opamp2 (Fig. 16) could be used to adjust the



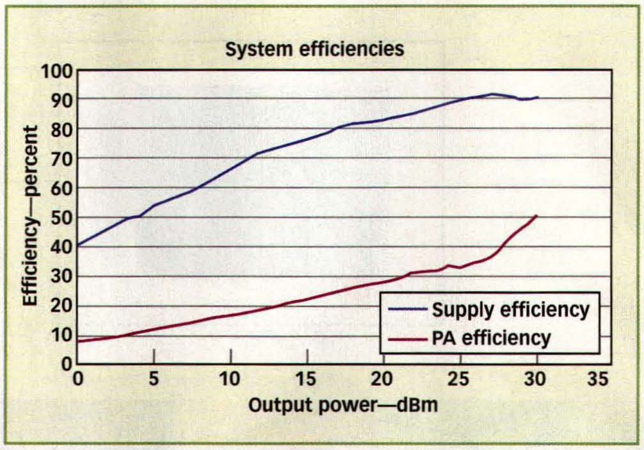
17. RF5117 performance with the V_{cc} and V_{reg} adjusted through P_{out} is illustrated here.



18. This shows RF5117 linearity performance with V_{cc} and V_{reg} adjusted through P_{out}.



19. Control-loop voltages for V_{cc} PA and V_{reg} adjusted through P_{out} are shown here.



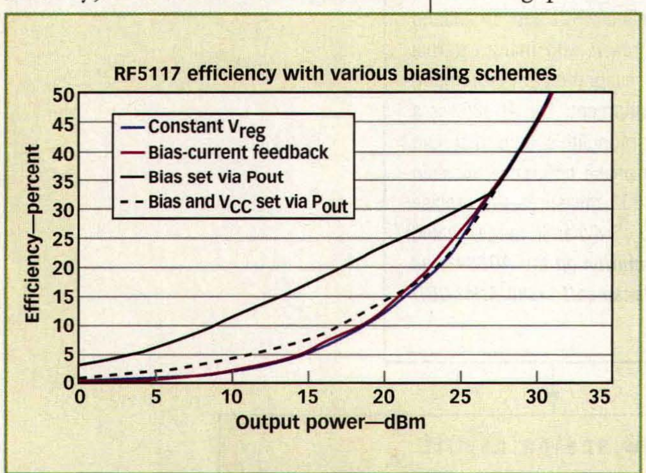
20. System efficiencies when V_{cc} PA and V_{reg} are adjusted via P_{out} can be seen here.

point where the DC-to-DC converter provides the maximum output voltage. The bias control could then be readjusted through changes in the 30-kΩ (voltage-to-current converter) resistor.

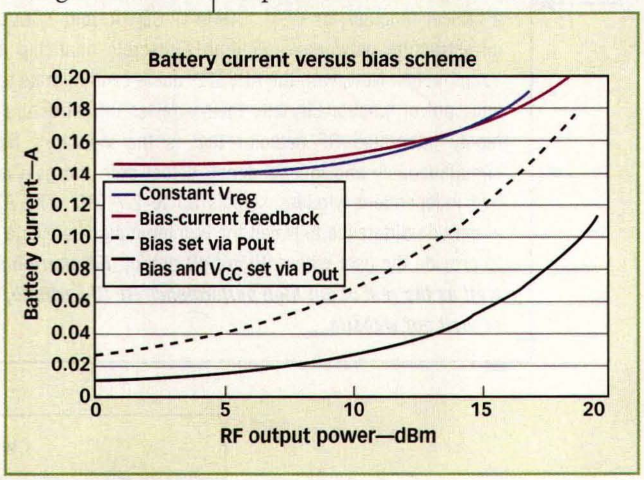
Monitoring the voltage and current out of the DC-to-DC converter provided an opportunity to study the losses in the system. **Figure 20** shows the calculated efficiencies based on this data. The supply efficiency is computed from the power out of the power supply (battery) to the power into the PA. This also includes the V_{reg} bias current and the power dissipated in the control chip. Essentially, the DC-to-DC converter with added bias losses determines this efficiency. The PA efficiency is determined by the DC power into the RF devices and the RF output power. Unfortunately, as a result of the hardware configuration, the PA bias current is included in the supply-efficiency category. The effect of the HBT knee voltage causes the reduction in the PA efficiency at low output powers. At 0-dBm output power, the applied voltage across the RF devices is slightly less than +0.6 VDC and the knee voltage is approximately +0.4 VDC. The difference between the applied voltage and the knee voltage is sufficient for linear-amplifier operation, as evidenced by the ACPR measurements (Fig. 18). The knee voltage degrades the efficiency. At higher output powers, however, the HBT knee voltage becomes a smaller fraction of the applied voltage and efficiency improves.

In all but the second RF PA biasing approach (with the bias-current feedback), the V_{reg} voltage is determined from a bandgap reference voltage. The automatic adjustment of the bias current could also be implemented using the bias-current feedback approach, and the results should be equally satisfying as the effect of temperature and process variations in the PA would be reduced.

Figure 21 shows a comparison of the total RF efficiencies for the four biasing approaches. Plotting all the data on the same axis allows direct comparisons between the biasing schemes presented in this article. The first two methods with constant V_{reg} and the bias-current feedback were configured to provide approximately the same bias current at low RF powers so little difference would be expected in the efficiencies. Adjusting the bias current as a function of the output power provided obvious improvement below +20 dBm. The efficiency improvement is limited to the reduc-



21. This shows total RF efficiency for the four biasing schemes.



22. Total current consumption for the four biasing schemes is illustrated here.

tions in bias current since the applied voltage is the same in both cases. When the voltage applied to the PA is varied through the DC-to-DC converter, the big gain in efficiency results from the transformer-like action that trades voltage for current (i.e., power out equals power in). At 0 dBm, the relative improvement exceeds a factor of 10.

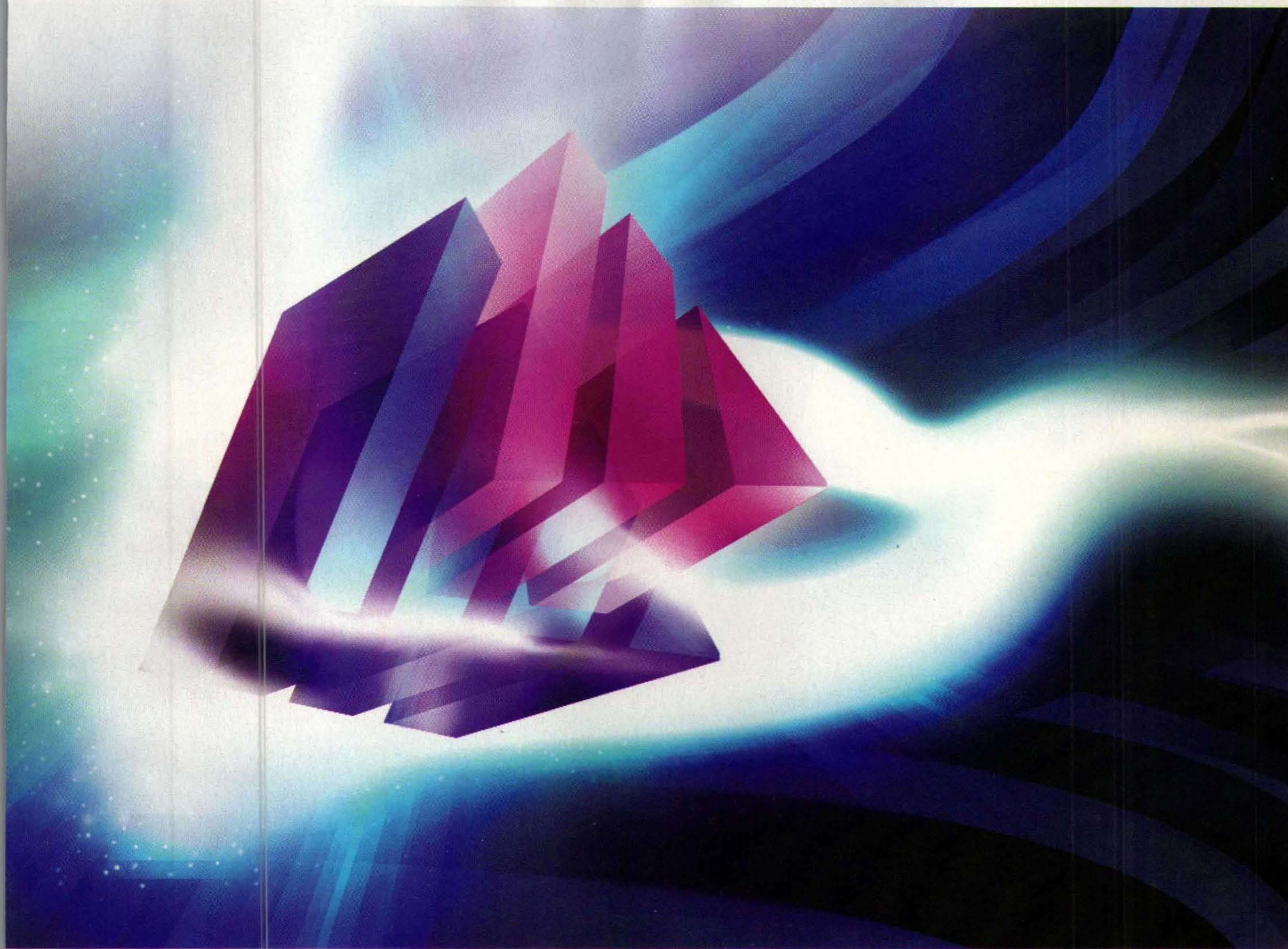
As was seen with the DC-to-DC converter, an efficiency number in itself can be misleading. The key parameter is reduced current consumption for a particular output power. **Figure 22** shows a comparison of the total current requirements for the four biasing methods at low RF output powers. Adjusting the PA bias current and voltage as a function of the output power yields the best results. This approach provides remarkable reductions in current requirements. The total quiescent current in this case, which includes the bias chip, the DC-to-DC converter, and the RF5117, is below 8 mA.

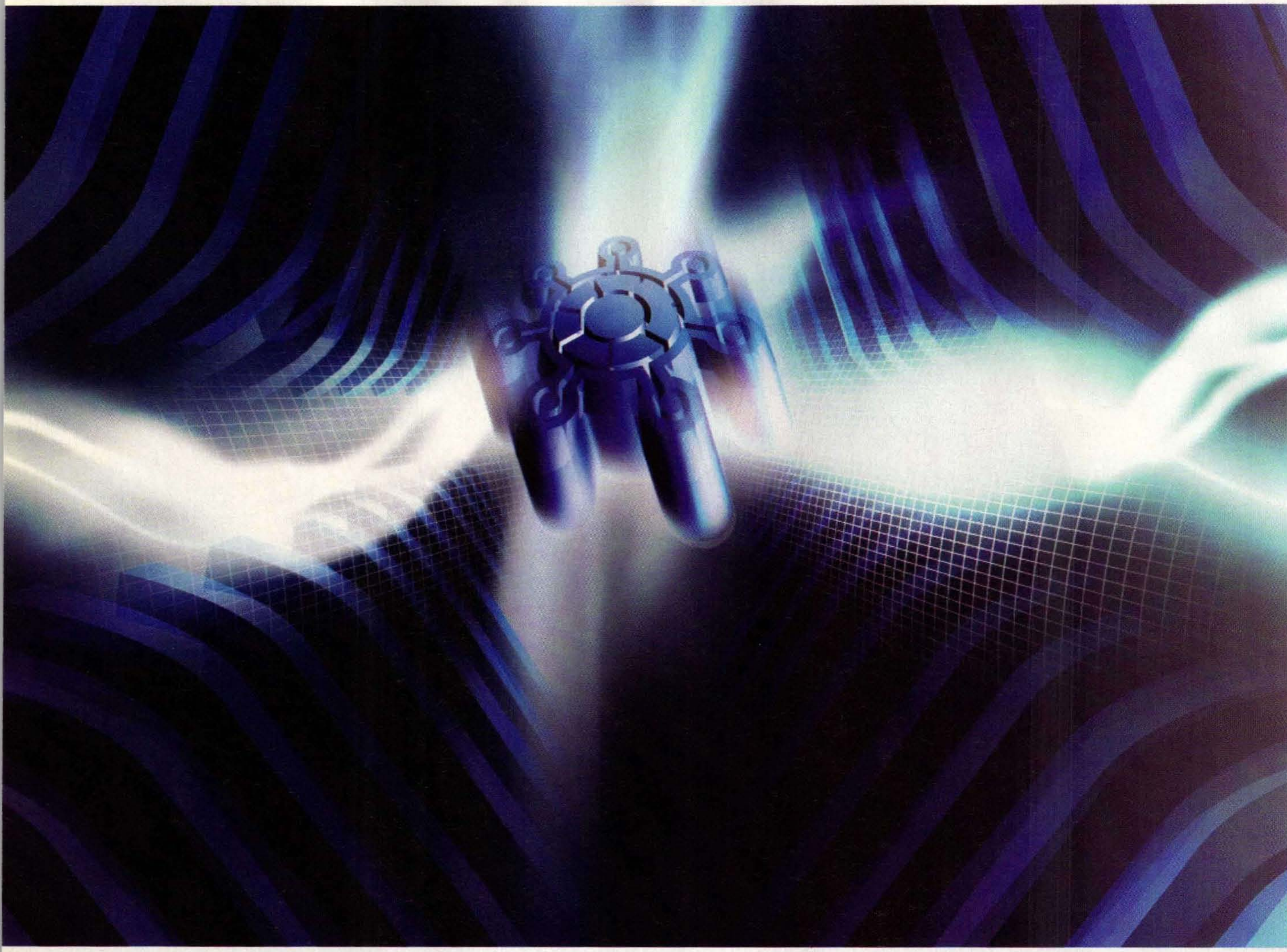
These results should be interpreted based on cost and system requirements. For example, if a system requires +27-dBm output power 90 percent of the time, the simplest (constant V_{reg} or bias-current feedback) biasing scheme may be most inexpensive. But if the system requires +10-dBm or less output power most of the time, then one of the bias-reduction schemes may be best. The least-expensive solution is to not use the DC-to-DC converter. However, if reducing the battery drain is of prime importance, the cost of the DC-to-DC converter may be negligible in the overall operating scheme.

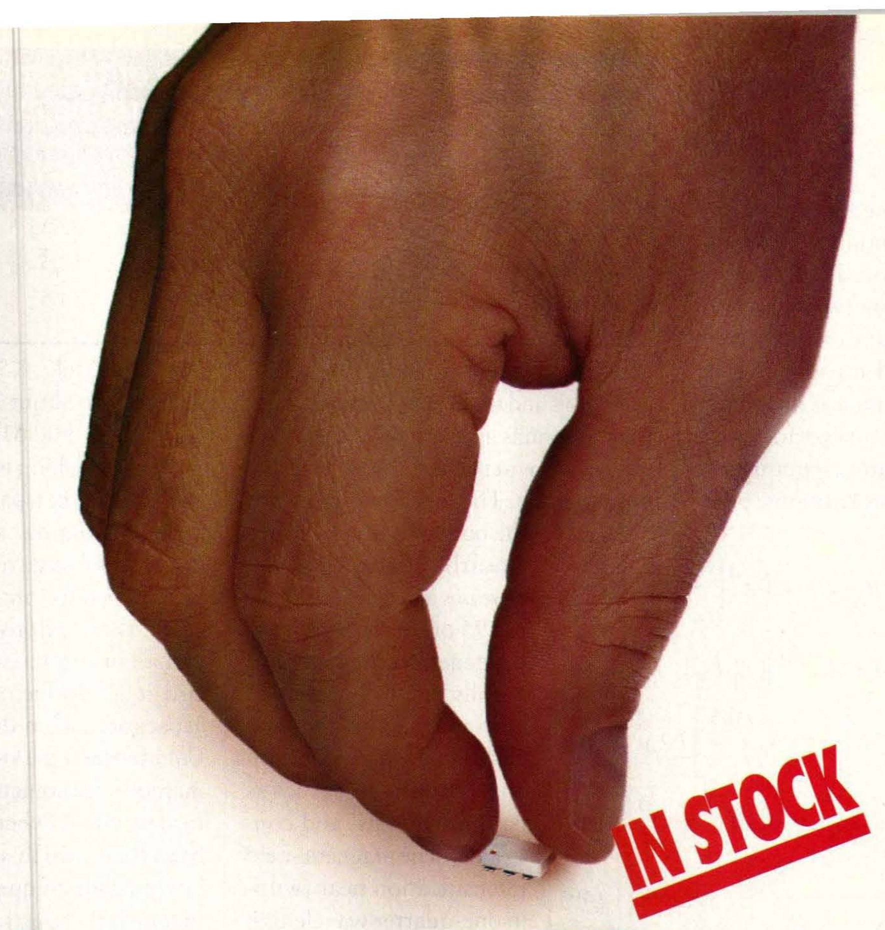
The gain expansion that occurs when the bias current is changed should not cause a problem. System requirements for flat gain typically result from an open-loop approach used to control the output power. The only requirement from a system that uses a feedback approach for power control is that the power is monotonic. The benefits of varying the PA bias point should outweigh a desire for flat gain, especially when the control loop provides a measure of the output power as a byproduct of its operation. MRF

ACKNOWLEDGMENT

The author would like to thank Jeff Potts of RF Micro Devices for the samples of the DC-to-DC converter that were used in the variable V_{CC} measurements.







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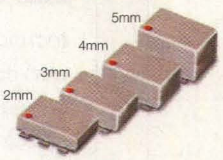

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ADEX-10L	+4	10-1000	7.2	60	16	3	2.95
ADE-1	+7	0.5-500	5.0	55	15	4	1.99▲
ADE-1ASK	+7	2-600	5.3	50	16	3	3.95
ADE-2ASK	+7	1-1000	5.4	45	12	3	4.25
ADE-6	+7	0.05-250	4.6	40	10	5	4.95
ADEX-10	+7	10-1000	6.8	60	16	3	2.95
ADE-12	+7	50-1000	7.0	35	17	2	2.95
ADE-4	+7	200-1000	6.8	53	15	3	4.25
ADE-14	+7	800-1000	7.4	32	17	2	3.25
ADE-901	+7	800-1000	5.9	32	13	3	2.95
ADE-5	+7	5-1500	6.6	40	15	3	3.45
ADE-5X	+7	5-1500	6.2	33	8	3	2.95
ADE-13	+7	50-1600	8.1	40	11	2	3.10
ADE-11X	+7	10-2000	7.1	36	9	3	1.99▲
ADE-20	+7	1500-2000	5.4	31	14	3	4.95
ADE-18	+7	1700-2500	4.9	27	10	3	3.45
ADE-3GL	+7	2100-2600	6.0	34	17	2	4.95
ADE-3G	+7	2300-2700	5.6	36	13	3	3.45
ADE-28	+7	1500-2800	5.1	30	8	3	5.95
ADE-30	+7	200-3000	4.5	35	14	3	6.95
ADE-32	+7	2500-3200	5.4	29	15	3	6.95
ADE-35	+7	1600-3500	6.3	25	11	3	4.95
ADE-18W	+7	1750-3500	5.4	33	11	3	3.95
ADE-30W	+7	300-4000	6.8	35	12	3	8.95
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ADE-1MH	+13	2-500	5.2	50	17	3	5.95
ADE-1MHW	+13	0.5-600	5.2	53	17	4	6.45
ADE-10MH	+13	800-1000	7.0	34	26	4	6.95
ADE-12MH	+13	10-1200	6.3	45	22	3	6.45
ADE-25MH	+13	5-2500	6.9	34	18	3	6.95
ADE-35MH	+13	5-3500	6.9	33	18	3	9.95
ADE-42MH	+13	5-4200	7.5	29	17	3	14.95
ADE-1H	+17	0.5-500	5.3	52	23	4	4.95
ADE-1HW	+17	5-750	6.0	48	26	3	6.45
ADEX-10H	+17	10-1000	7.0	55	22	3	3.45
ADE-10H	+17	400-1000	7.0	39	30	3	7.95
ADE-12H	+17	500-1200	6.7	34	28	3	8.95
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ADE-20H	+17	1500-2000	5.2	29	24	3	8.95

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Introducing Loop Antennas For Short-Range Radios

Part 5 of this article series addresses the basics of loop-antenna design for short-range radios with integrated PLL Txs, comparing several matching methods for efficiency and performance.

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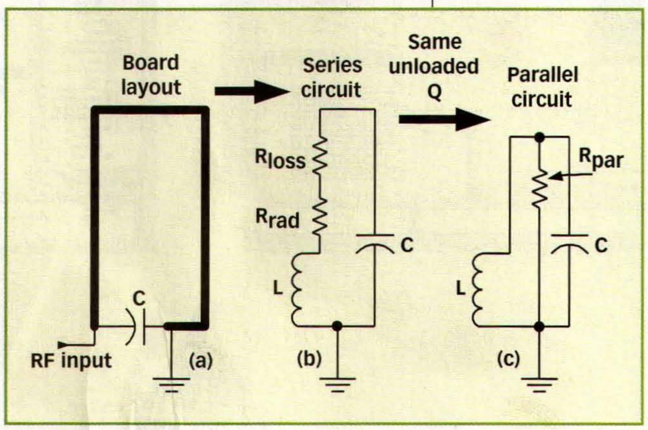
Short-range-radios are only as strong as their weakest component links, including the antenna. Previously, Parts 1, 2, 3, and 4 of this series (see *Microwaves & RF,* September and October 2001 and February and March 2002, respectively) covered one-way short-range system design, including link budgeting, regulatory issues, and some issues of silicon (Si) design at the transmit side. Part 5 of this article series now addresses the basics of loop-antenna design for short-range radios with integrated phase-locked-loop (PLL) transmitters (Txs), comparing the unmatched and the tapped-capacitor matching methods for efficiency and performance. Part 6 will cover the printed-transformer matched-loop antenna and understanding differential drive. The concluding Part 7 will cover physical-board design and practical regulatory compliance.

Printed-loop antennas are commonly used with unlicensed short-range radios, due to requirements for small size, ruggedness, and low cost. The frequency range is generally 285 to 470 MHz (see Part 2 of this series), where a full-sized quarter-wave whip antenna measures 6.28 to 10.4 in. (16.0 to 26.3 cm) in length. The large size usually eliminates consideration of whip antennas for these applications, resulting in the acceptance of printed-circuit-board (PCB) antennas as the most popular solution. PCB antennas generally exhibit only 1-to-20-percent radiation efficiency, but are small, easy to design (with the exception of significant errors in some published matching methods), insensitive to design errors since they



11. The physical implementation of a loop antenna (a) is shown next to its standard model (b) and an impedance transformed model (c) where all resistances are viewed in parallel.

Table 7: Calculated performance of the unmatched and matched 12 × 34-mm loop antenna at 434 MHz.

	Mismatch loss (dB)	Total efficiency	2nd harmonic rejection (dB)	3rd harmonic rejection (dB)
Unmatched	11.3	0.59 percent	22.1	23.6
Tapped capacitor	0	8 percent	50.6	52.0
Transformer	0	8 percent	41.5	36.0

usually must be tuned anyway, and provide a modest amount of harmonic suppression (improved by matching, as discussed later). However, since most of the available literature on loop-antenna matching is aimed at paging-receiver (Rx) applications, there is little information on the harmonic performance needed to meet regulatory requirements in transmit mode. This harmonic problem shall receive considerable attention in this and the next two articles. Also, loop antennas actually have enhanced efficiency when positioned near the human body. The low conductivity of the human body decreases a nearby electric field and increases a magnetic field (see p. 295 of ref. 8), leading to the general view that electrically small "magnetic" loop antennas are the most efficient for equipment worn on the human body, such as pagers, RF tags, and controllers. The magnetic-field intensification near (within one-quarter wavelength of) the human body is approximately 4.5 dB at 285 MHz, dropping to about 2.8 dB at 470 MHz and 0 dB at 900 MHz.

$$R_{rad} = 320\pi^4\left(\frac{A^2}{\lambda^4}\right) \quad (45)$$

$$R_{rad} = \left(3.84\times10^{-30}\right)\left(L_1 L_2\right)^2 f^2 \quad (46)$$

$$R_{loss} = \left[l\left(\pi f \mu_0\right)^{0.5}\right] / 2w \quad (47)$$

$$R_{loss} = \left[\left(L_1 + L_2\right) / w\right]\left(2.61\times10^{-7}\right)(f)^{0.5} \quad (48)$$

$$\eta_r = \frac{R_{rad}}{R_{rad} + R_{lossL} + R_{lossC}} \quad (49)$$

In use, the loop inductance is usually considered to be a parallel resonance with a variable tuning capacitor so that the driver sees a large real load which must be matched for optimum power delivery. Other options to manual tuning include using resistors to modify the circuit quality factor (Q) to allow fixed capacitors and on-die automatic tuning. Unfortunately, the losses imposed by these methods are sometimes unacceptable. In particular, when a low-cost wideband Rx is used that prevents setting the intermediate-frequency (IF) bandwidth to match the spectral occupancy of the transmitted signal, then "averaging"

as described in Part 2 is often used to maintain link quality. This averaging requires higher radiation efficiency and, thus, usually a well-matched and individually tuned high-Q loop antenna.

Figure 11 shows a standard loop-antenna model where series loss resistances are moved to provide a total parallel equivalent resistance that maintains the same Q in the loop. A matched single-ended driver would provide similar loading by driving into the nongrounded end of the capacitor, and Q will be cut approximately in half from the limit set by radiation and loss resistance. If the loop is directly driven by a lower-impedance power amplifier (PA) [unmatched], then Q will be lower still.

The radiation resistance of a loop, under the condition that it is electrically small (perimeter less than 0.3 λ), is provided as Eq. 45,[9]
where:

A = loop area (perimeter of inside edge of trace) in square meters and

λ = wavelength in meters.

For the frequencies and sizes normally used, this equation generally holds out to approximately the second to fourth harmonic and is adequate to use in predicting the lower-order harmonic performance where regulatory compliance is more commonly an issue. At higher frequencies where the antenna is not electrically small, the current in the antenna varies as a function of position, and must be taken account of as outlined in ref. 11 or through simulation. For a rectangular antenna with sides L_1 and L_2 fabricated on copper (Cu)-clad laminate, the given Cu conductivity of 5.8×10^7, Eq. 45 becomes Eq. 46.

An expression for loss resistance derived from fundamental principles (skin-depth-based analysis), assuming that line width is much greater than line thickness, but thickness is also much greater than skin depth (true for practical boards), is Eq. 47,
where:

l = the total perimeter of the antenna in meters, measured at the center of the trace;

w = the width of the trace in meters;

σ = conductivity; and

μ = permeability.

For the common rectangular antenna case with Cu trace and with permeability of 1.256×10^{-6}, Eq. 47 becomes Eq. 48.

The radiation efficiency of the loop is commonly provided as Eq. 49.

For a particular driving current to the loop, this expression follows immediately from power being i^2R. An alert reader may immediately wonder about driving current changing with variation in loss and matching resistance if a perfect match is provided by other circuitry. A

simple analysis can show that if match is maintained, the same expression results if efficiency is defined as the radiated power divided by the total driving power. Though often neglected, losses associated with the resonating capacitor are usually significant and are counted in the denominator of Eq. 49 as another series resistance-loss term. Good COG capacitors will typically have series loss resistances of 0.1 to 0.2 Ω, variable capacitors series loss resistances from 0.1 to 0.5 Ω, and X7R and Z5U dielectrics series loss resistances of 0.5 and 1 Ω (visit the Murata website at www.murata.com for an excellent database of these losses over capacitor construction, value, and frequency). These capacitor losses can dramatically affect radiation efficiency and matching, and can have a moderate effect on harmonics.

It is often helpful in analysis to transform losses between series and parallel modes, which is valid around a narrow range of frequency. Using series losses as the base mode, we may define:

$$Q_S = \frac{X_S}{R_S} \qquad (50)$$

from which analysis the following highly useful set of basic relations is Eqs. 51-56:

$$R_P = R_S\left(Q_S^2 + 1\right) \qquad (51)$$

[SEE EQS. 52 TO 56 ABOVE]

$$X_P = X_S\left(\frac{Q_S^2 + 1}{Q_S^2}\right) \qquad (52)$$

$$L_P = L_S\left(\frac{Q_S^2 + 1}{Q_S^2}\right) \approx L_S\left(\textit{for}\text{ high Q}\right) \qquad (53)$$

$$C_P = C_S\left(\frac{Q_S^2}{Q_S^2 + 1}\right) \approx C_S\left(\textit{for}\text{ high Q}\right) \qquad (54)$$

$$R_P = R_S\left(1 + Q_S^2\right) = \frac{\left(\omega L_S\right)^2}{R_S} + R_S \approx \frac{\left(\omega L_S\right)^2}{R_S}\left(\textit{for}\text{ high Q}\right) \qquad (55)$$

$$R_P = R_S\left(1 + Q_S^2\right) = \frac{1}{\left(\omega C_S\right)^2 R_S} + R_S \approx \frac{1}{\left(\omega C_S\right)^2 R_S}\left(\textit{for}\text{ high Q}\right) \qquad (56)$$

$$L = \frac{\mu}{2\pi} l\ln\left(\frac{8A}{lw}\right) \qquad (57)$$

$$R_{LStot} = R_{rad} + R_{lossL} + \left(\omega_h^2 LC\right)^2 R_{lossC} \qquad (58)$$

Of course, to achieve a loop resonance, an expression is required for loop inductance. A remarkably simple formula for inductance of a polygon of general shape that is usually good to within 5 percent is provided by ref. 10 as Eq. 57,

where:

l = the perimeter as measured at the inside edge of the trace,

w = the width, and

A = the area.

Consider a numerical example for a loop that will later be matched using several other approaches. Assume operation at 434 MHz (a common European choice) with a rectangular antenna measuring 3.4 × 1.2 cm on the inside, with trace width of 2 mm, and with a capacitor with series loss at this frequency of 0.138 Ω. The loss resistance can be calculated as 0.250 Ω, the radiation resistance as 0.0227 Ω, the total series resistance as 0.286 Ω, and the resulting maximum efficiency as 7.95 percent. From Eq. 57, the inductance is 52.9 nH and the resonating capacitance is thus 2.54 pF. The unloaded Q is 505 and the equivalent parallel resistance is 72.9 kΩ.

Drivers on low-power Txs would normally have an output impedance of from 50 Ω to several thousand Ω, so a direct connection across this loop is obviously a bad mismatch that would not provide the maximum possible efficiency. The low impedance of the typical driver would also lower the Q drastically and reduce the harmonic rejection of the antenna. Despite these disadvantages, an unmatched loop is occasionally used in these applications, so the analysis is provided as follows. The total loss resistances of the loop antenna (where losses are modeled as a resistance is series with the inductor) are shown in Eq. 58,

where:

R_{rad} = radiation resistance,

R_{lossL} = ohmic loss resistance in the loop, and

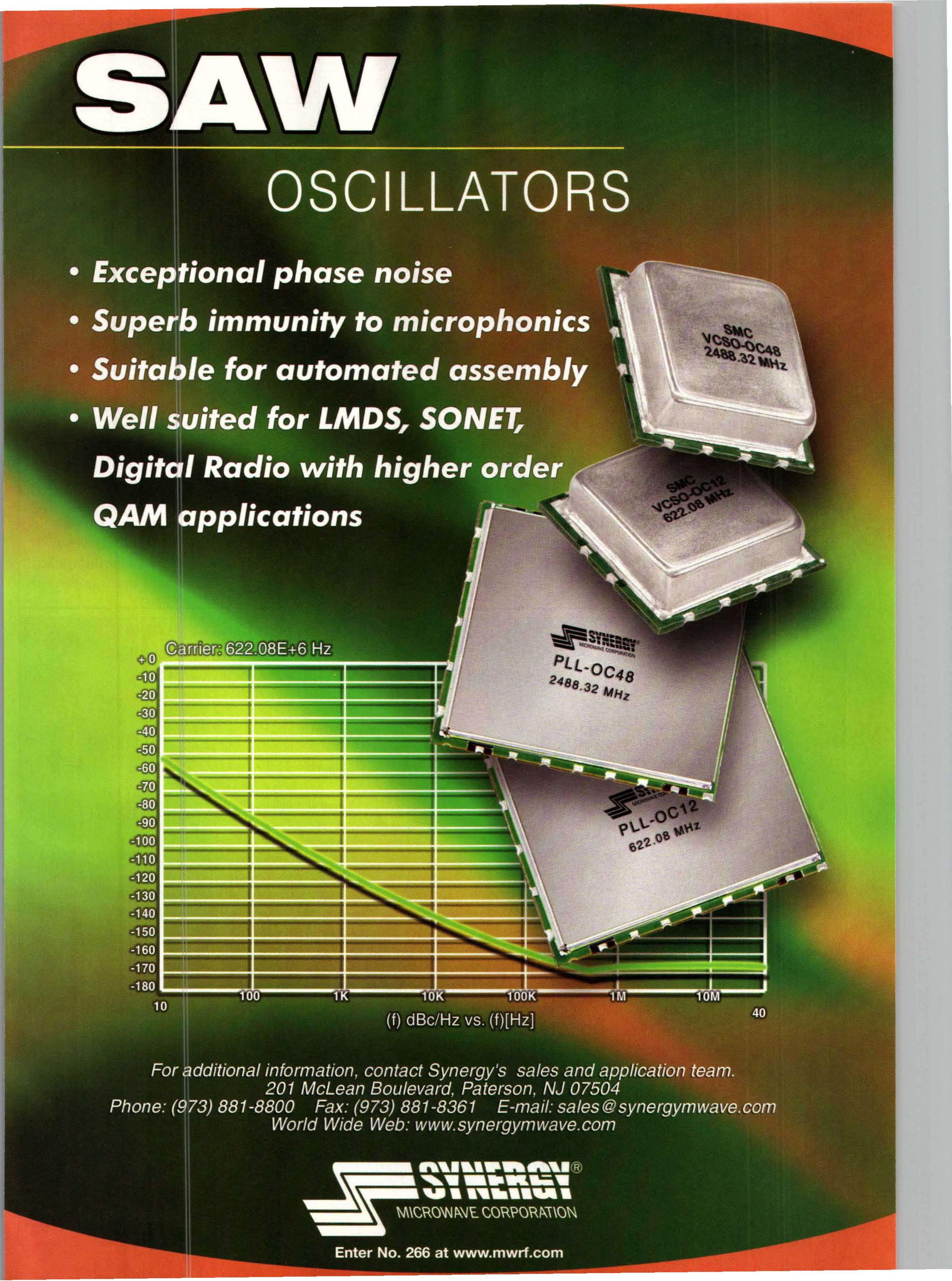
R_{lossC} = capacitor series loss resistance.

The coefficient of R_{lossC} is 1 at the fundamental frequency, but greater than 1 at the harmonic frequencies. This coefficient results from moving the capacitor series loss, R_{cs}, to be in series with the inductor for modeling purposes. This sum may be represented in parallel form at the fundamental and harmonic frequencies by Eq. 55, yielding a quantity here known as R_{PtotH}, where H represents the harmonic number and is 1 for the fundamental frequency. Assuming the antenna still satisfies the constant spacial-current approximation for the first few harmonics, the radiation efficiency for the fundamental and first few harmonics can be written as:

$$\eta_H = \frac{R_{rad}}{R_{LStot}} \qquad (59)$$

where it is understood that R_{rad} must be found from Eq. 45 at the appropriate harmonic H. Converting impedances to admittances, we may write a handy current-divider function expressing the fraction of the total current available from the driver at each harmonic frequency that flows through this parallel resistance at each harmonic frequency and is thus radiated. Defining G_{PtotH} as the total parallel admittance at each harmonic H (where H = 1 at the fundamental), this divider function is provided by Eq. 60.

In Eq. 60, the term Y_{driver} is used because at the harmonic frequencies the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$$D_{IH} = Mag\left(\frac{G_{PtotH}}{G_{PtotH} + G_{driver} + j\left[\omega_h C - \frac{1}{\omega_h L}\right]}\right) \tag{60}$$

$$\frac{P_H}{P_1} \approx \frac{\eta_H}{\eta_1}\,\frac{0.316\left(D_{IH}\right)^2 R_{PtotH}}{\left(D_{I1}\right)^2 R_{Ptot1}} \tag{61}$$

$$MisMatchLoss = \frac{4\left(\frac{Z_{in}}{R_{driver}}\right)}{\left(\frac{Z_{in}}{R_{driver}}\right)^2 + \frac{2Z_{in}}{R_{driver}} + 1} \tag{62}$$

driver impedance and admittance are not normally purely real. At the fundamental frequency, the circuit is resonant and the imaginary component is zero, but at the harmonics it is dominated by the capacitance and most of the driver current available at the harmonics is shunted to ground and does not radiate. The ratio of each harmonic current to driver fundamental current is needed to determine the harmonic rejection. For approximation purposes, however, we may assume that the fundamental power is 10 dB over the first few harmonics (typical for a compressed Class A single-ended PA), but that the antenna is 5 dB more directional for the harmonic frequencies. The harmonic rejection (radiated harmonic to carrier power) in measured field strength for each harmonic H may, thus, be approximated to about ±5 dB accuracy as Eq. 61.

The mismatch of the directly driven loop (power applied at the loop capacitor) is large. In general, for a source with impedance R_{driver} driving a load of parallel impedance Z_{in}, the "mismatch loss" (which does not include efficiency losses) may be determined by Eq. 62.

Table 7 provides example performance numbers for the example loop antenna (the 1.2 × 3.4-cm loop operating at 434 MHz) when directly driven by a source of 1.4 kΩ impedance. The mismatch loss in this case is approximately 11 dB, the efficiency about 8 percent (resulting total efficiency of less than 1 percent), and the harmonic rejection is just over 20 dB. With this example, there is risk of failing harmonic regulatory requirements (see Part 2 of this article series) in addition to generally weak link performance, although the poor Q of the antenna may eliminate the need for tuning. This table also includes a line for the transformer-matched loop antenna to be presented in Part 6.

The large mismatch and relatively poor harmonic suppression of the unmatched loop antenna may be much improved

by the tapped capacitor matching method (shown in single-ended form in **Fig. 12**). Here, the fundamental definition of matching is seen in elegant simplicity, where matching means viewing the total set of loss impedances in the loop antenna as a single input parallel impedance $R_{par} = Z_{in}$ that yields the same unloaded Q. When R_{par} matches the driver resistance, the maximum power-transfer theorem is satisfied and the loaded Q will be one-half of the unloaded Q. Intuitively, the tapping may be seen to yield a down-impedance transform through conservation of energy, with the voltage at the tap point lowered from the inductor voltage by the capacitive divider action, and thus requiring a lower impedance (if all loss resistance is modeled at that point to provide the same Q) at the tap point to dissipate the same power. Pursuing this analytically will yield the result shown in Eq. 63 for parallel R_{in} as a function of inductor parallel resistance R_p.

To develop design equations for the tapped-capacitor case and to understand its harmonic performance, the broadband admittance looking into the capacitor tap is written as Eq. 64, where:

R_s = the resistance in series with the inductor that models all losses.

It is desirable to solve this equation for C_1 and C_2 to force the desired R_{in} and resonant frequency. The reciprocal of the real part of Eq. 64 yields the input resistance at resonance and provides one equation. Setting the imaginary part equal to zero at the desired resonant frequency yields the other. The results are shown in Eqs. 65 and 66.

Equation 64 also provides the way to understand the harmonic performance of the tapped-capacitor loop antenna. For the large impedance transform from parallel resistance across the inductor to parallel R_{in} at the tap point C_2 will normally be much larger than C_1, and much larger than the C of the unmatched loop. Thus, C_2 dominates the input admittance at the tap and shunts most current to ground, greatly improving harmonic rejection. This

$$Z_{in} = \left(\frac{1}{1 + \frac{C_2}{C_1}} \right)^2 R_p \qquad (63)\ (\textit{applies at resonance})$$

$$G_{in} = \frac{R_s}{R_s^2 + \left(\omega L - \frac{1}{\omega C_1} \right)^2} + j \left(\omega C_2 - \frac{\omega L - \frac{1}{\omega C_1}}{R_s^2 + \left[\omega L - \frac{1}{\omega C_1} \right]^2} \right) \qquad (64)$$

may be quantified in a manner similar to the unmatched loop, where the "current-divider function" for harmonic current that flows in the real part of the input admittance (where it must flow to be radiated) is shown in Eq. 67.

Despite this divider function, some current still flows in the real part of loop-radiation resistance that is transformed to the input and it is this current that radiates power. The radiated power at the fundamental (H = 1) and at each harmonic (where the loop is still "small") is illustrated in Eq. 68, where:

i_{rmsH} = the root-mean-square (RMS) current available from the source at harmonic frequency H*Fundamental, and is the fundamental current when H = 1 (where $D_{IH} = 0.5$ due to the match

$$C_1 = \frac{1}{\omega_0\left(\omega_0 L - \sqrt{Z_{in} R_s - R_s^2}\right)} \quad (65)$$

$$C_2 = \frac{L - \frac{1}{\omega_0^2 C_1}}{R_s^2 + \left(\omega_0 L - \frac{1}{\omega_0 C_1}\right)^2} \quad (66)$$

$$D_{IH} = Mag\left(\frac{\mathrm{Re}\left[G_{inH}\right]}{G_{inH} + G_{driver}}\right) \quad (67)$$

$$P_{radH} = \frac{\eta_H\left(i_{rmsH}\,D_{IH}\right)^2}{\mathrm{Re}\left(G_{inH}\right)} \quad (68)$$

condition).

The harmonic rejection relative to the carrier is provided by the ratio of harmonic power in Eq. 68 to the radiated carrier power (also from Eq. 68 with H=1) degraded by the extra directivity of the antenna at the harmonic frequency. Assuming the applied harmonics are 10 dB down from the carrier and that the antenna is no more than 5 dB more directive for the harmonics yields the approximation:

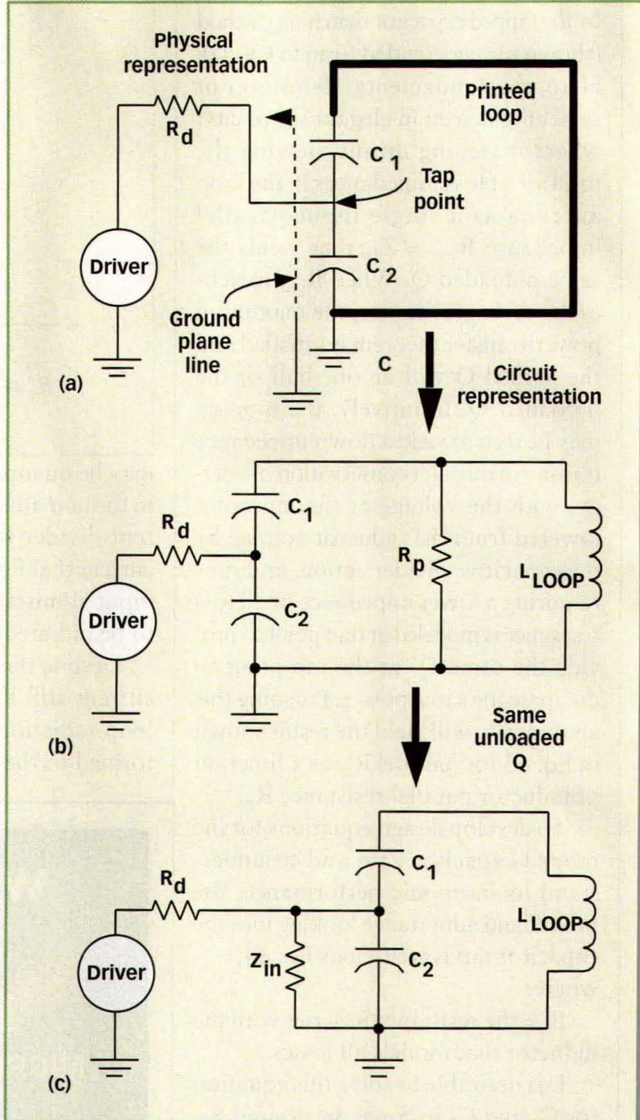
$$\frac{P_H}{P_1} \approx \frac{\eta_H}{\eta_1}\,\frac{1.26\left(D_{IH}\right)^2}{\mathrm{Re}\left(G_{inH}\right) Z_{in1}} \quad (69)$$

At harmonic frequencies, Eq. 64 can be simplified to:

$$G_{inH} \approx \frac{R_{sH}}{\left(\omega_H \mathrm{L}\right)^2} + j\omega_H C_2 \quad (70)$$

Technically, C_1 and C_2 must be well-controlled to meet desired resonance and input impedance conditions. In practice, with a variable capacitor for C_1 or C_2, the tapped-capacitor method can yield a good match, near perfect resonance, and more than 40-dB harmonic rejection. For the example loop antenna at 434 MHz, with a driver impedance of 1.4 kΩ, L = 52.9 nH, C_1 = 2.95 pF, and C_2 = 18.3 pF, the predicted second harmonic levels are –50.6 dBc, and the predicted third harmonic levels are –52 dBc. There is little mismatch loss, so the total efficiency is the loop and capacitor efficiency of approximately 8 percent. These harmonics will normally pass all regulatory requirements, but to achieve such low loop harmonic levels, a PCB designer must beware of parasitic radiation from traces and bond wires that may actually dominate measured performance.

The material presented in this article should allow first-order understanding and design of the unmatched and tapped-capacitor loop antennas. The relations shown allow approximate prediction of radiated power and harmonics, at least over the first few harmonics where the loop is still electrically small. For higher harmonics where the loop is not electrically small, use of an electromagnetic (EM) simulator is recommended. The tapped-capacitor antenna has been found to be capable of excellent harmonic suppression, so together with its higher efficiency due to good matching, it is an excellent choice. However, the transformer-loop antenna to be presented next month can provide equal efficiency and acceptable harmonic suppression with lower parts count. There has been some incorrect information published on the operation of the transformer loop antenna, so methods based on the underlying EMs that are fundamentally sound will be shown. Next month's Part 6 will also cover understanding the use of differential drive on all these antenna types. The concluding Part 7 will deal with practical issues such as basic regulatory measurements, nonideal harmonic radiation from integrated-circuit (IC) pins and supply lines, and cost trade-offs in controlling these effects. MRF



12. These three circuit representations illustrate single-ended tapped-capacitor-antenna matching.

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Explore the future of backward-compatible WLANs

NEW HIGH-PERFORMANCE WIRELESS local-area networks (WLANs) must interoperate with active 802.11b equipment so that upgrade paths from 2.4-GHz systems are easily identified. Some think that a dual-band solution that addresses 802.11a and 802.11b concerns is the answer. A three-page application note, "High Speed WLANs: Moving Toward a 5-GHz Solution" by Mark Roberts, vice president of marketing for IceFyre Semiconductor Corp. (Kanata, Ontario, Canada) tries to answer this question and more by analyzing the current state of the backward-compatible WLAN market and attempting to predict its future.

In the note, Roberts crunches some numbers and shows that, while the currently installed numbers of 802.11b equipment is high (16.5 million clients by the end of 2003), it is less than the 1996 numbers for 10BaseT Ethernet (120 million clients). Roberts notes that a large percentage of the WLAN marketplace has no use for backward-compatible systems, and the industry should focus on this portion of the market.

Along the way, Roberts addresses the dual-mode market, discusses an optimal dual-band solution, and explores the limitations of alternative architectures. He surmises that, to be successful, a high-performance WLAN solution must exist in an uncrowded band to ensure high data throughput and offer sufficient channels to support pico-cell deployment in high-user-density areas.

Roberts concludes that the backward-compatible dual-band solution market is only a small percentage of the total available WLAN market. Since the active 802.11b marketplace being relatively small, developers must decide how to reach out to this niche market while at the same time moving toward a large-scale high-performance WLAN marketplace penetration. An optimized backward-compatible solution will be a great stepping stone to scaling these heights, but the optimized 802.11a core will remain the most important part of getting there. This application note is available as a free download from the company's website.

IceFyre Semiconductor Corp., 300-411 Legget Dr., Kanata, Ontario K2K 3C9, Canada; (613) 599-3000, FAX: (613) 599-4965, Internet: www.IceFyre.com.

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Combining basic microwave measurement practices with techniques targeting higher frequencies can yield optimum millimeter-wave measurements.

Improve millimeter-wave spectrum measurements

IDEAL SPECTRUM MEASURING is difficult at high frequencies and millimeter-wave regions present their own sets of problems. However, combining basic microwave measurement practices with techniques targeting higher frequencies can yield optimum millimeter-wave measurements with little trouble. According to an application note entitled "Improve Your Millimeter-Wave Spectrum Measurements" by Ben Zarlingo of Agilent Technologies (Santa Clara, CA), even inexperienced designers can bypass common errors and make successful measurements. Common errors in practice will still generate poor measurements despite using the the best equipment if an designer does not follow basic steps.

The term "millimeter wave" refers to frequencies of 26.5 or 30.0 GHz and beyond. This is the point at which wavelengths fall below 10 mm and the terminology then changes from centimeters to millimeters. The techniques described in the application note cover millimeter-wave measurements in coaxial environments to approximately 50 GHz.

Modern spectrum analyzers cover frequencies from intermediate frequency (IF) and RF through millimeter wave, and are perfect tools for designers in this range. Improving spectrum measurements can be accomplished if the designer uses cables/adapters specific to millimeter-wave applications; uses a connector saver on the analyzer; uses a torque wrench on all connections; uses the same set of cables, connectors, and routing; performs calibrations or alignments often, especially when conditions change; understands the measurement plane and its characteristics; connects the analyzer to the circuit using the simplest and straightest path possible; and reviews instrument specifications to comprehend performance level. Following these tips can greatly enhance measurements. This application note is available as a free download from the company's website.

Agilent Measurement Solutions, P.O. Box 3828, Englewood, CO 80155-3828; (800) 452-4844 ext. 7349, Internet: www.agilent.com/find/mentor.

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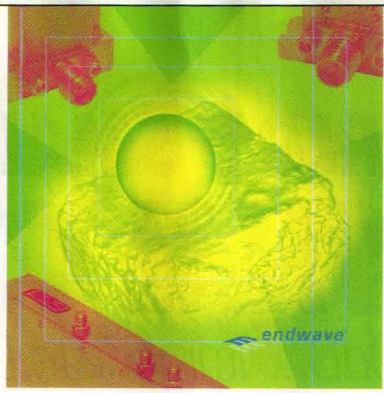
cover story



YIGs Tune High-Speed Millimeter-Wave Radios

Permanent-magnet YIG oscillator technology is alive and well and fueling a line of frequency synthesizers for high-data-rate digital microwave radios.

Yttrium-iron-garnet (YIG) oscillators enjoy a rich tradition in broadband military systems and microwave designs, including signal generators and spectrum analyzers. But these tunable sources are also critical components in modern commercial millimeter-wave radios. The low phase noise of these oscillators enables these radios to achieve high spectral efficiency over wide tuning ranges while meeting demanding size and cost requirements.

MARINUS (RON) KORBER
Sr. Scientist

DAN TEUTHORN
Director of Engineering

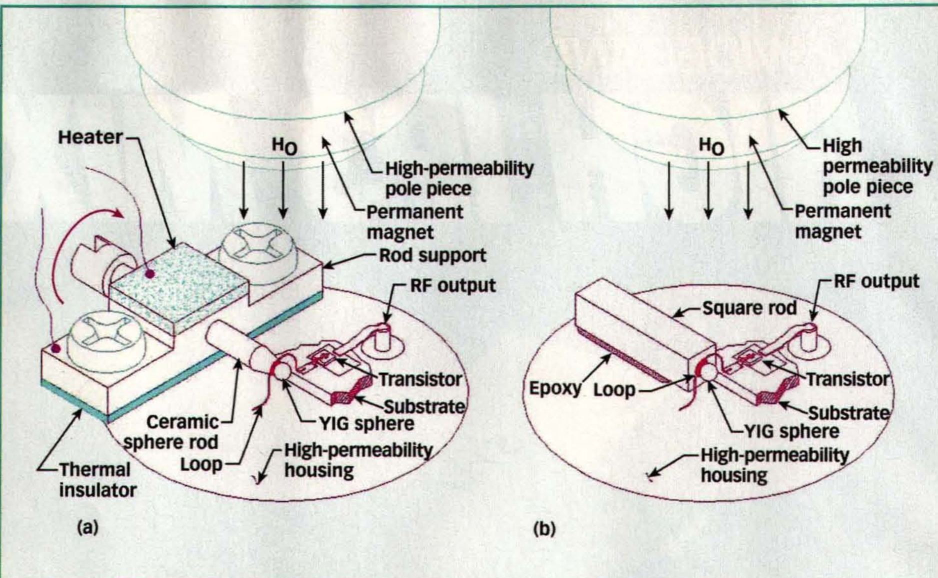
DOUG LOCKIE
Founder and Executive Vice President, Advanced Marketing

Endwave Corp., 990 Almanor Ave., Sunnyvale, CA 94085; (408) 522-3100, FAX: (408) 522-3102, Internet: www.endwave.com.

The size and cost of YIG oscillators have continued to decrease to the point where high-performance permanent-magnet YIG oscillators are less than a 0.60-in. (1.52 cm) on a side at less than $150.00 each in large quantities. As increasing demands for higher data rates push the complexity of the modulation schemes in digital radios, the phase-noise performance of YIG oscillators makes them viable candidates as the local-oscillator (LO) source in these radios. In particular, permanent-magnet YIG oscillators such as the MicroYIG and MiniYIG sources from Endwave (Sunnyvale, CA) provide tuning ranges as wide as 1 GHz at center frequencies ranging from 3 to 11 GHz.

The MicroYIG and MiniYIG oscillators can be specified with center frequencies from 3 to 11 GHz, the former with tuning ranges to ±500 MHz and the latter with tuning ranges to ±1000 MHz. The smaller MicroYIG sources measure 1.0 × 0.4 × 0.7 in. (2.54 × 1.02 × 1.78 cm), while the larger MiniYIG sources measure 1.0 × 1.0 × 0.8 in. (2.54 × 2.54 × 2.03 cm).

A YIG oscillator relies on the behavior of a highly polished YIG sphere in a magnetic field. A highly polished YIG sphere (with typical diameter of 14 mils) behaves as a high-quality-factor (high-Q) tuning element within a magnetic field.



1. Traditional YIG sources require a large number of parts and labor-intensive assembly (a). The Endwave approach eliminates parts, labor, and cost, as well as allowing automated assembly (b).

Magnetic tuning is possible since the resonant frequency of a YIG sphere is a linear function of the magnetic field where the sphere is placed. A YIG sphere will tune at a constant rate of 2.8 MHz/G of applied field. Using a small coupling loop, a narrow band of microwave energy can be coupled into and extracted from the YIG sphere. The coupling loop and resonator behave much like a parallel resonant circuit with intrinsic resistor (R), inductor (L), and capacitor (C) values that are dependent upon sphere properties such as the unloaded Q and the loop geometry. When used with a negative-resistance source (such as an unstable transistor), a tunable low-phase-noise oscillator results.

In a MicroYIG oscillator, the YIG sphere rests within an airgap between the permanent magnet and a high-permeability, metal-injection-molded (MIM) housing on which the oscillator substrate rests. The housing and its cover serves as a flux guide for the permanent magnet's magnetic field and a mu-metal shield to decouple the YIG sphere from external electromagnetic (EM) fields.

The permanent magnet provides the magnetic field needed to saturate the YIG sphere. The resonant frequency of the YIG sphere is tuned through the main tuning coil around the magnet, by decreasing or increasing the strength of the magnetic field. A second tuning coil is used for phase-locking the YIG oscillator, as well as providing frequency modulation (FM) with bandwidths in excess of 400 kHz.

The uniformity of the magnetic field around the YIG sphere is critical. If the field is not uniform, parts of the sphere will oscillate at slightly different frequencies, resulting in degraded phase-noise performance. Permanent magnets inherently produce somewhat nonuniform fields which can be smoothed out by attaching a small ferromagnetic disk—known as a field straightener—to the face of the permanent magnet.

Since MIM housings have higher electrical resistivity than machined metal housings, a higher main-coil modulation bandwidth is possible with these MIM YIG oscillators. For example, the main-coil modulation bandwidth of the MicroYIG series oscillators is better than 35 kHz, typically wide enough for many phase-locking applications. The main-coil tuning sensitivity (or modulation sensitivity) of these MicroYIG oscillators is typically 5 MHz/mA, with virtually constant modulation sensitivity across the tuning band.

An important aspect of permanent-magnet YIG oscillator design is the control of frequency drift as a function of temperature. In the traditional approach to assembling a YIG oscillator, epoxy is used to attach a YIG sphere to the end of a ceramic rod. The rod is held by a rod support that allows it and the sphere, which is positioned inside of a coupling loop, to be rotated during an initial alignment process **(Fig 1a)**. The YIG sphere, which acts similar to a cubic crystal, is mounted on the rod face in a particular plane orientation (the <110> plane) lying parallel to the magnetic field. Four different zero-drift angles lie in this plane and the sphere must be rotated close to one of these to minimize frequency drift as a function of temperature. Although not common to commercial designs, a small heater is often employed in military YIG oscillators to achieve demanding frequency-drift specifications. The heater requires an additional current source. The rod support, the heater chip, the ceramic sphere rod, and the time to assemble these parts and rotate the sphere to a correct axis add manufacturing cost to the oscillator assembly of a conventional YIG oscillator.

To improve upon tradition, the engineers at Endwave developed a new patented approach to YIG oscillator design **(Fig. 1b)**. In this approach, the YIG sphere is fastened to the face of an inexpensive square, plastic, injection-molded rod. The YIG sphere is epoxied to the rod face in a position that is preset to a zero-temperature axis **(Fig. 2)**. By using this approach, the formerly labor-intensive YIG sphere/rod alignment process has been transformed into a "pick-and-place" step in a more-automated oscillator-assembly process.

Another detail to consider in YIG oscillator design is the effects of vibration. Endwave's engineers typically work with customers, encouraging them to use a shock-mount kit **(Fig. 3)** supplied with demonstration units to learn more about shock-mounting techniques.

YIG sources are well-known for their application in electronic-warfare (EW) and other military systems. But YIG oscillators are also vital to the operation of high-data-rate digital microwave radios, especially for those employing highly efficient modulation schemes, including 256-state quadrature amplitude modulation (256QAM). A YIG source with wide tuning range can increase the number of channels possible

Continued on page 100

PRODUCT *technology*

VNA Series Gains Range And Features

These instruments fill two important ranges of frequency coverage while simplifying low-frequency device characterization and making calibrations easier and faster.

JACK BROWNE
Publisher/Editor

Improved performance was part of the original launch of the PNA series vector-network analyzers (VNAs) by Agilent Technologies (Santa Rosa, CA) two years ago. With time comes additional models, including the 45-MHz-to-20-GHz model E862A and the 45-MHz-to-40-GHz model E8363A, as well as enhanced features such as calibration refinements and logarithmic sweep capabilities. The new two-port, four-receiver (Rx) VNAs complement Agilent's existing 3-, 6-, and 9-GHz two- and three-port analyzers, as well as the 50-GHz PNA Series VNAs. A 67-GHz PNA model is scheduled for introduction later this summer.

With these additions to the PNA series of VNAs **(see table)**, the instrument family now includes instruments designed to serve a wide array of wireless, aerospace, and defense applications. The PNA series analyzers employ an architecture based on Microsoft Windows 2000, rather than a proprietary instrument operating system similar to its predecessors (and nearly every other network analyzer). The instruments also employ a mixer-based signal-separation approach, rather than the more-traditional sampler technology. Sampler-based Rxs can deliver exceptional performance, but they suffer from relatively high conversion loss.

The mixer-based approach allows the instruments to achieve exceptionally broad dynamic range, while also delivering measurement speed as fast as 26 µs per point. As the flagship of the microwave PNA series, the E8364A (introduced in September 2001) maintained this level of performance to 50 GHz, arguably the fastest, widest-dynamic-range VNA at that frequency. The E8362A and E8363A

The E8362A and E8363A VNAs at a glance

Measurement frequency range	45 MHz to 20 GHz (E8262A) 45 MHz to 40 GHz (E8363A)
Number of ports	2
Measurement receivers	4 Hz
Maximum output power	+3 dBm at 20 GHz –4 dBm at 40 GHz
Dynamic range at test port (when equipped with option 014)	136 dB at 20 GHz 119 dB at 40 GHz
Trace noise (1 kHz IF BW)	Less than 0.006 dB (0.1 deg RMS)
Measurement speed (1601 pts)	7 updates/s (E8362) 4 updates/s (E8363A)
High-stability timebase	Standard
Extended power range	Available
Bias tees	Available
Time-domain capability	Available

retain this performance at maximum frequencies of 20 and 40 GHz, respectively **(see figure)**. This level of performance will be extended shortly with the 67-GHz model E8361A.

Logarithmic-sweep capability is one of the most important enhancements delivered to the PNA series with the introduction of the E8362A and E8363A. Log-sweep capability complements the instruments' existing linear, power, segment, and continuous-wave (CW) sweep capabilities, and makes analysis of broadband measurements much easier. Log-sweep capability provides benefits in several measurement situations. A linear sweep tends to compress lower-frequency information into a smaller portion of the overall display. In contrast, the log-sweep function shows the response within each decade with similar horizontal display resolution. This provides a more-detailed view of a component's response at low and high frequencies.

The PNA series of VNAs maintains 26-μs performance at maximum frequencies of 20 and 40 GHz, respectively.

Log sweep provides good resolution at low frequencies without requiring the higher frequencies to be measured with the same high resolution. For example, log sweep provides a Bode plot display where the slope of the response is indicative of the number of poles in the filter (one pole provides 20 dB/decade, two poles 40 dB/decade, etc.), a format with which most designers are familiar.

The PNA Series VNAs have also been enhanced with new calibration capabilities, all of which are designed to make the calibration process simpler and faster. Two enhancements in particular are geared toward the 3-, 6-, and 9-GHz instruments and users who employ Agilent's electronic-calibration (ECal) modules. ECal is an electronic system that provides full one-, two-, three-, or four-port calibration with a single connection. It significantly reduces the chance of operator error, which is always possible when mechanical calibration standards are used, since mechanical calibration requires connecting numerous standards to each measurement port.

The four-port ECal capability introduced in revision 2.5 firmware allows all two or three ports of the RF PNA models to be recognized by the ECal module, so that they may be used whenever an N-port calibration with a four-port ECal module is desired. This allows the user to fully reap the benefits of ECal by making the fewest possible connections. Another ECal enhancement provides users who have 8509xC ECal modules with the ability to embed the effects of adapters, cables, or fixtures in the characterization, extending one or more of the module's calibration planes to the ends of the connectors on these devices.

To minimize calibration errors, several new features have been incorporated within the error-reporting system of the PNA series VNAs. For example, errors are now displayed on the VNA screen. A pop-up error-display window now includes a help-identification number to find additional information within the help system. The calibration wizard within each PNA instrument includes error messages that inform users of an error, along with possible solutions. The E8362A and E8363A are priced from $73,900 and $95,000, respectively, and are currently shipping. **Agilent Technologies, Test and Measurement Organization, 5302 Stevens Creek Blvd., MS 54LAK, Santa Clara, CA 95052; (800) 452-4844, Internet: www.agilent.com/find/pna.**

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ICs Send 100 Mb/s Using UWB Technology

The promise of UWB technology has been realized in the Trinity chip set, four ICs capable of wirelessly transmitting 100 Mb/s while consuming very low power levels.

JACK BROWNE
Publisher/Editor

Ultrawideband (UWB) technology has intrigued wireless designers for several years with its theoretical potential for large information bandwidths at low transmit power levels. XtremeSpectrum (Vienna, VA) has become the first company to put that theory into practice, with the Trinity chip set, four integrated circuits (ICs) capable of wirelessly transferring data at 100 Mb/s using less than 200 mW of power.

The low-cost chip set is suitable for wireless data transfers between multimedia products, such as digital cameras and camcorders and personal computers (PCs).

The Trinity chip set is the only commercial wireless solution capable of transmitting two or more high-definition (HD) streams simultaneously. The chip set consists of the model XSI 141 medium-access-control (MAC) IC (based on the emerging IEEE 802.15.3 standard), the XSI 122 baseband controller, the XSI 112 UWB transceiver, and the XSI 102 low-noise amplifier (LNA). The LNA is fabricated with silicon germanium (SiGe). The remaining ICs are fabricated with a standard 0.18-µm digital silicon (Si) complementary-metal-oxide-semiconductor (CMOS) process. One of the benefits of the UWB technology is that the performance (data rate) of these chips will get better with improvements in semiconductor process technology. No changes need to be made to the basic IC architectures.

UWB technology has been called the "carrierless" communications format, since information is carried by a series of short pulses rather than by a modulated carrier. In contrast to conventional modulated carriers which occupy a relatively narrow band of frequencies, such as the 2400 to 2483 MHz of Bluetooth, UWB signals are essentially narrow pulses that occupy a relatively wide portion of frequency spectrum. According to the mandate of the US Federal Communications Commission (FCC), however, the energy of these low-power pulses must be concentrated within the band from 3.1 to 10.6 GHz with minimal out-of-band emissions to interfere with existing systems, such as the Global Positioning System (GPS), satellite radio, and personal communications services (PCS).

The UWB solution offered by XtremeSolution, a company founded in 1998, employs biphase UWB transmissions. This approach uses two different-phase pulses to represent digital ones and zeros in a transmission. Biphase UWB approaches can send pulses right side up or upside down, with the orientation determining whether a pulse

Continued on page 100

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is a digital 1 or a digital 0. Since biphase techniques can send a large number of pulses in a short time period, longer coding sequences can be added to a signal than with monopulse techniques. This enables a receiver to lock onto the longer code, eliminating the effects of interference and multipath fading and filtering out unwanted noise.

The biphase UWB approach implemented in the Trinity chip set inherently yields twice the overall power efficiency of monopulse UWB transmissions with twice the data rate as a function of distance. The company employs spectrally shaped pulses to reduce out-of-band emissions and comply with the FCC's requirements for unlicensed UWB transmissions.

Prior to the Trinity chip set, the biphase UWB modulation approach had not been employed in commercial designs due to the complexity of the chip architecture and the speed limitations of the semiconductor processes capable of providing the level of integration needed. But with the availability of high-speed Si CMOS and SiGe processes that also lend themselves to high levels of integration complexity, the signal-processing architectures needed to implement UWB biphase modulation could be manufactured with high-enough yield to support a low-cost chip set.

According to XtremeSpectrum's CEO, Martin Rofheart, "the desire for wireless connectivity between consumer electronic devices has been great, but there hasn't been a technology that could simultaneously deliver the consumer market's three critical criteria: low price, low power, and high data rate." He notes that the Trinity chip set embodies a technology that can provide consumers with the performance needed: "Trinity, as the name implies, is the first chip to do so, and will enable a new class of consumer electronics products to deliver 'wirelike' video quality using a wireless medium."

The Trinity chip set is supported by a compact-form-factor reference design, available as part of the company's model XSI-EVK-101 evaluation kit. The reference design features a patented omnidirectional antenna that is fabricated on standard printed-circuit-board (PCB) material and can be manufactured in high volumes at low cost. The antenna, of course, is vital to meeting the US FCC's spectral mask for unlicensed UWB transmissions, and for achieving the full high-date-rate performance capability of the chip set. P&A: $19.95 (four-chip set, 100,000 qty.) and $50,000 (model XSI-EVK-101 evaluation kit). **XtremeSpectrum, Inc., 8133 Leesburg Pike, Suite 700, Vienna, VA 22182; (703) 269-3000, Internet: www.xtremespectrum.com.**

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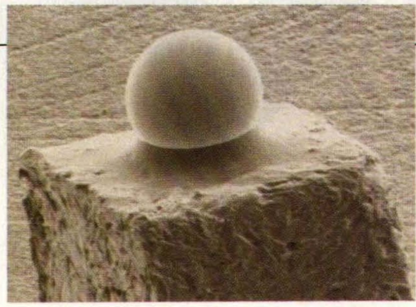
cover story *(continued)*



2. By fastening a YIG sphere to the face of an inexpensive square, plastic, injection-molded rod, the position of the YIG resonator can be preset in Endwave oscillators to eliminate tuning time and expense.

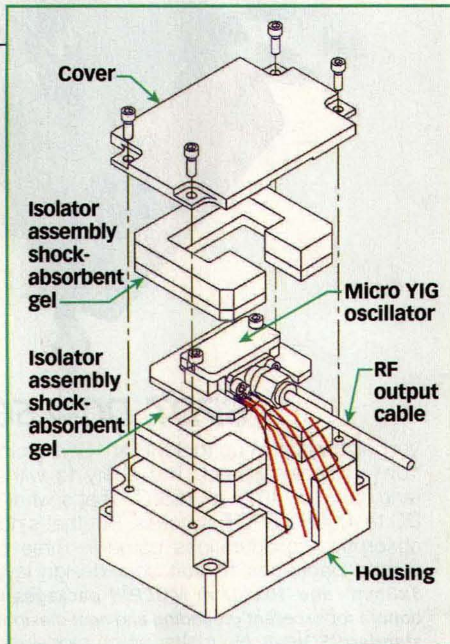
Continued from page 94

with a microwave radio, and when frequency-stabilized within a practical synthesizer design, can provide the low-phase-noise performance required to support complex modulation formats.

For example, Endwave's Series 50 frequency synthesizers can be implemented with either MicroYIG or MiniYIG oscillators, allowing cost-versus-bandwidth trade-offs within the same package. The synthesizers incorporate a single-loop, fractional-N phase-locked-loop (PLL) architecture for low phase noise over a wide tuning range. As examples, models SYN-50A-00, SYN-50B-00, and SYN-50C-00 feature operating ranges of 4.5 to 7.0 GHz, 7 to 10 GHz, and 10 to 12 GHz, respectively, with minimum tuning steps of 125 kHz. The phase noise is typically –128 dBc/Hz offset 100 kHz from any carrier and –143 dBc/Hz offset 1 MHz from any carrier.

With a 2-kHz loop bandwidth, Series 50 synthesizers can achieve step sizes as small as 250 kHz at 10 GHz while maintaining nonharmonic spurious performance of –50 dBc or better. Reference spurs occur at 62.5-kHz offsets due to the fixed prescaler.

By using a unique shock-mounting approach in the MicroYIG and MiniYIG oscillators, as well as in the Series 50 synthesizer housings, phase hits are minimized. Carefully selected materials help achieve good mechanical dampening



3. This shock-mounting concept helps minimize the effects of vibration.

from –35 to +75°C). **Endwave Corp., 990 Almanor Ave., Sunnyvale, CA 94085; (408) 522-3100, FAX: (408) 522-3102, Internet: www.endwave.com.**

Enter No. 51 at www.mwrf.com

SiC MESFET Drives 10-W Output Power

MODEL CRF-22010 IS a silicon carbide (SiC) power metal semiconductor field-effect transistor (MESFET) designed for use in Class A and Class AB amplifiers. The device, which is usable to 3 GHz, delivers at least 10-W output power (1-dB compression) at 2 GHz with typical output power of 12 W. The drain efficiency is typically 45 percent and the gain is typically 12 dB. Designed for +48-VDC systems, the transistor is supplied in either a metal-ceramic flange-style housing or drop-in package.

Cree, Inc., 4600 Silicon Dr., Durham, NC 27703; (919) 313-5300, FAX: (919) 313-5451, e-mail: sales@cree.com, Internet: www.cree.com.

● **Enter No. 64 at www.mwrf.com**

Compact Amplifier Hits High Intercept Point

MODEL ULA 808-XX IS a miniature-packaged amplifier designed for applications from 1800 to 2100 MHz. The amplifier provides 14 dB typical linear gain with +28-dBm typical output power. The third-order intercept point (IP3) is typically +48 dBm, with a minimum of +46 dBm. For lower-frequency applications, the model ULA 818-XX amplifier operates from 800 to 1000 MHz with slightly higher gain and similar dynamic range.

Microwave Technology, Inc., 4268 Solar Way, Fremont, CA 94538; (510) 651-6700, FAX: (510) 651-2208, e-mail: info@mwtinc.com, Internet: www.mwtinc.com.

● **Enter No. 65 at www.mwrf.com**

Dual-Synthesizer IC Serves WCDMA Designs

MODEL IBM7012 IS a wideband-code-division-multiple-access (WCDMA) dual-synthesizer integrated circuit (IC) that can handle transmit and receive chores. Consisting of integrated voltage-controlled-oscillator (VCO) and fractional-N synthesizer functions, the IC offers frequency ranges of 3250 to 3450 MHz and 4170 to 4400 MHz. The synthesizer phase noise is –140 dBc/Hz offset 10 kHz from the carrier, while spurious levels are no more than –70 dBc and typically as good as –90 dBc. The synthesizer IC works with input frequencies from 900 to 1200 MHz.

IBM Microelectronics Div., 1580 Route 52, Building 504, Hopewell Junction, NY; Internet: www.ibm.com/chips/support/howtobuy.html.

● **Enter No. 66 at www.mwrf.com**

Internally Matched FET Yields 180-W Power

MODEL PTF 102022 IS an internally matched (to 50 Ω) silicon field-effect transistor (FET) designed for wideband-code-division-multiple-access (WCDMA) applications from 2110 to 2170 MHz. The transistor provides 12.5-dB linear gain and delivers 27-W average power when operating with two WCDMA carriers. Under continuous-wave (CW) conditions, the transistor provides 180-W output power at 1-dB compression with 11.5-dB gain and 45-percent efficiency.

Ericsson Microelectronics, 18275 Serene Dr., Morgan Hill, CA 95037; (877) GOLD-MOS, e-mail: rfpower@ericsson.com, Internet: www.ericsson.com/rfpower.

● **Enter No. 67 at www.mwrf.com**

Power FET Pushes 160-W CW Output

MODEL PTF 102005 IS an internally matched (to 50 Ω) silicon (Si) field-effect transistor (FET) designed for wideband-code-division-multiple-access (WCDMA) applications from 2110 to 2170 MHz. The transistor provides 13-dB linear gain and delivers 30-W average power when operating with two WCDMA carriers. Under CW conditions, the transistor provides 160-W output power at 1-dB compression with 13-dB linear gain and 45-percent efficiency. The gold (Au)-metallized device, which can withstand mismatches equivalent to a 10:1 VSWR, is provided in a flanged metal-ceramic housing.

Ericsson Microelectronics, 18275 Serene Dr., Morgan Hill, CA 95037; (877) GOLD-MOS, e-mail: rfpower@ericsson.com, Internet: www.ericsson.com/rfpower.

● **Enter No. 68 at www.mwrf.com**

Low-Noise MMIC Amplifier Handles 24 to 33 GHz

MODEL ES/FMM5703VZ IS a low-noise-amplifier (LNA) monolithic-microwave integrated circuit (MMIC) designed for applications from 24 to 3 GHz. The noise figure is only 2.5 dB (typical) at 32 GHz, with associated gain of 17 dB at that frequency. The LNA provides +8.5-dBm output power at 1-dB compression with input/output (I/O) return loss of typically 10 dB. The unit is input- and output-impedance matched to 50 Ω.

Fujitsu Compound Semiconductor, Inc., 2355 Zanker Rd., San Jose, CA 95131-1138; (408) 232-9500, FAX: (408) 428-9111, Internet: www.fcsi.fujitsu.com.

● **Enter No. 69 at www.mwrf.com**

PA MMIC Drives 17.5 to 31.5 GHz

MODEL ES/FMM5804VY IS a power amplifier (PA) monolithic-microwave integrated circuit (MMIC) designed for use from 17.5 to 31.5 GHz. The 50-Ω impedance-matched MMIC provides +24.5-dBm typical output power over the frequency range, with minimum small-signal gain of 13 dB and typical gain of 17 dB. The power-added efficiency (PAE) is typically 10 percent. The input return loss is typically 15 dB, while the output return loss is typically 8 dB. The amplifier is supplied in a hermetic surface-mount (flip-chip) package.

Fujitsu Compound Semiconductor, Inc., 2355 Zanker Rd., San Jose, CA 95131-1138; (408) 232-9500, FAX: (408) 428-9111, Internet: www.fcsi.fujitsu.com.

● **Enter No. 70 at www.mwrf.com**

2 & 4WAY

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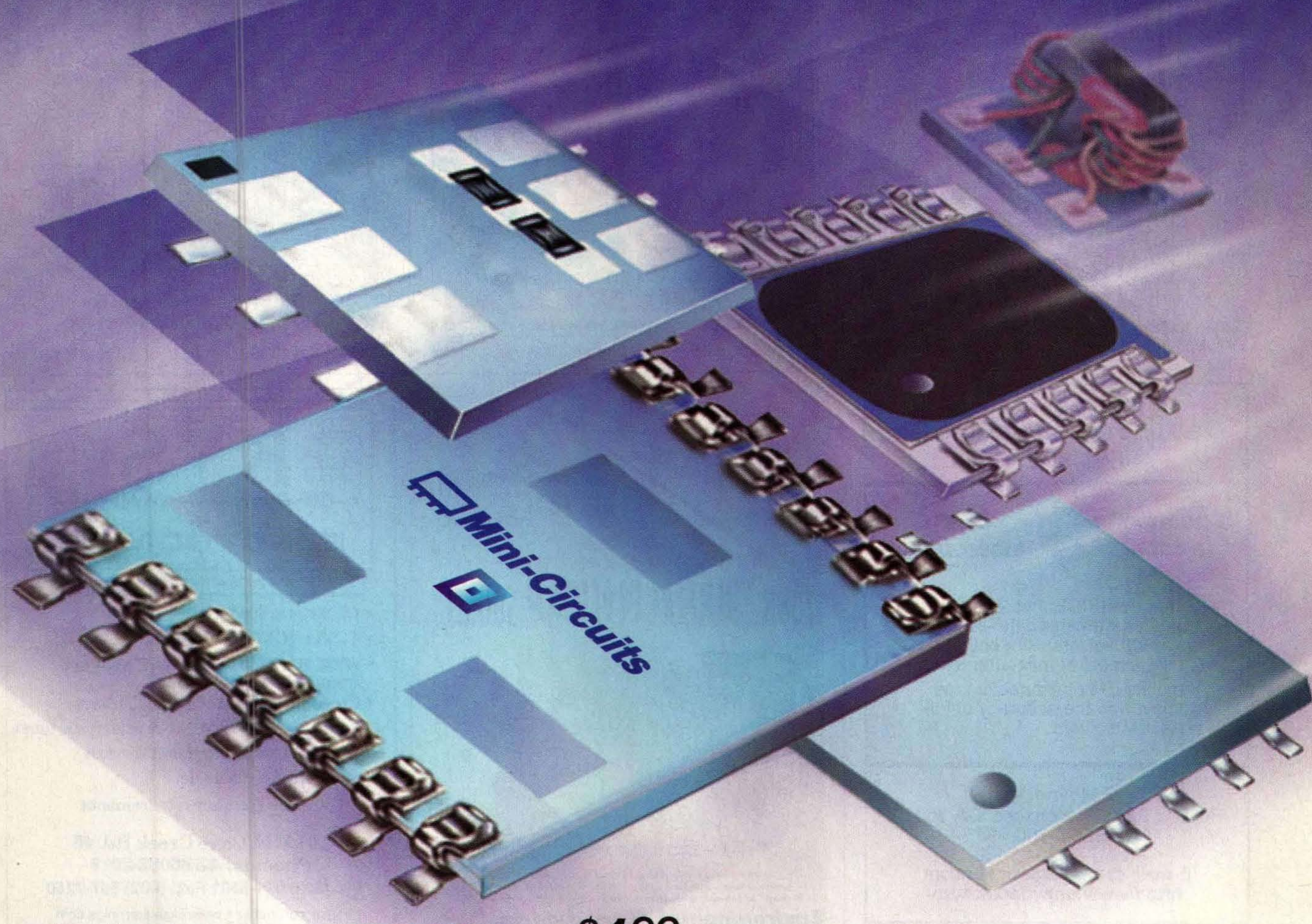

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2	0	SBA	4	1200-2600	16-22	0.4-0.8	5.0-10.0	6.95
2	0	SBB	5	800-2300	22-24	0.5-0.6	3.0-4.0	4.95
2	90	QBA	7	340-2400	21-28	0.25-0.80	3.0-7.0	6.95
4	0	SBD	1	1800-2600	20	1.0	8.0	9.95

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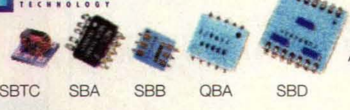

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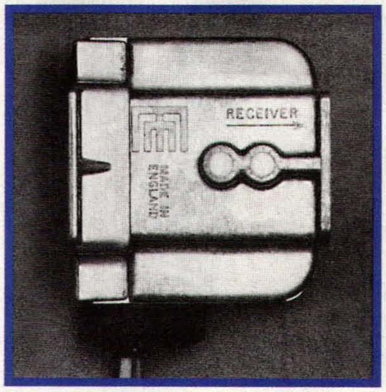

ALMOST 17 YEARS ago, English Electric Valve (EEV; Chelmsford, England) introduced a device known as the "dupletron," which integrated magnetron oscillators, duplexers, and solid-state receiver (Rx) electronics in a compact housing. The X-band device targeted low-cost marine radars.

next month

Microwaves & RF August Editorial Preview

Issue Theme: Wireless Applications

News

Near-record attendance at the Microwave Theory & Techniques Symposium (MTT-S) in Seattle, WA this past June was a sign that, although business in the high-frequency industry is tight, companies are still fighting and surviving, and often making significant technological advances along the way. The August issue will feature a special wrapup report on the recent MTT-S, highlighting some of the latest advances in hardware, software, and test equipment, with a sidebar story on the growing number of MEMS companies and products evident on the MTT-S exhibit floor.

Design Features

August will highlight activities in advanced mHEMT device processes and how these techniques may bring low-cost, high-frequency InP devices to the masses. Additional technical articles will offer two different approaches to Bluetooth testing, from the RF and baseband sides, while authors from Tality will show how to combine load- and source-pull techniques to improve the efficiency and linearity of PAs. Also, articles will detail a low-profile 5-GHz WLAN antenna, continue a series on the design of high-performance LNAs, and provide Part 6 of an article series on short-range radios.

Product Technology

The August Product Technology section will examine the emergence of the market for WLAN products, in particular a new integrated chip set with RF, IF, and baseband functions from a leading supplier of RF ICs. Additional product stories will unveil a wideband VNA system that works with modulated test signals, highlight a new supplier of LTCC circuit technology, examine several lines of narrowband and wideband frequency synthesizers with low phase noise and high resolution for digital microwave radio systems, explore the use of a specialized ASIC in high-performance TCXOs, and review a novel, low-cost chip set for HomePlug applications.